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- Control circuits for SMPS in tv receivers
- Image intensifiers — an introduction
- A simple hand-held night viewer

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Cover picture:
"Hello! My name's Brock. I'm a little worried about that long black thing pointing at me in the dark. I wonder what it is?" (Brock can find out why there is no privacy for badgers by reading the articles on image intensifiers in this issue.)
Photograph J.A. Bailey, courtesy Ardea, London.

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Control circuits for SMPS in tv receivers

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This article, the second in a series of four about television switched-mode power supplies, describes the requirements of the control circuits for tv SMPS. Two alternative integrated circuits described in detail are the TDA2581 and the TDA2640. The sync process on TDA2571 is also briefly described, along with its relationship to the rest of the tv SMPS circuit.

INTRODUCTION

A previous article in this series (Ref.1) examined the functions of the basic SMPS circuits which are most suited for use in tv receivers: the feasibility of using the SMPS circuit to provide the 150V supply for the horizontal deflection output stage was discussed and the possibility of also deriving the horizontal drive voltage from the SMPS was investigated. The control of such an SMPS is necessarily complex, and two integrated circuits have been developed, either of which may be used to control SMPSs in tv receivers. This present article discusses these two ICs (TDA2640 and TDA2581) and the control functions that they incorporate. The integrated sync processor (TDA2571) is also briefly described as this can be used in conjunction with the TDA2581.

REQUIREMENTS OF SMPS CONTROL CIRCUIT

To ensure reliable operation of a television receiver incorporating an SMPS, the SMPS output transistor must be correctly driven under all input (mains) and output (load) conditions. A control circuit must therefore be incorporated to perform the following functions:

1) provision of an ultrasonic switching waveform for the SMPS output transistor;
2) control of the duty factor (width modulation) of the SMPS output pulses to stabilise the supply voltage to the horizontal deflection circuit;
3) provision for synchronising the SMPS output pulses with the deflection waveform;
4) provision of a reference voltage for comparison with a voltage-dependent feedback from the output;
5) provision of an error amplifier for monitoring the gain and stability of the feedback loop;
6) limitation of the maximum duty factor of the SMPS output pulses in the event of the feedback voltage line being either short-circuited or disconnected at switch-on;
7) limitation of the inrush current when the receiver is switched on (slow-start);
8) overload and overvoltage protection;
9) protection against a low supply voltage to the IC.

A circuit capable of performing all these functions is rather complex and, if built from discrete components, it would be costly. We have therefore developed the integrated circuit TDA2640 to meet all of these requirements.

The SMPS may also be required to deliver base drive to the horizontal deflection output transistor. In this event, the control circuit must also incorporate phase control of the switching waveform so that the deflection circuit drive can be derived from a secondary winding on the SMPS output choke. To satisfy this requirement, an IC type TDA2581 has been developed.

The latest published data on these ICs should be consulted, and a further description of the TDA2640 is available in Refs.2 and 3.

The control loop

Fig.1 shows a control loop in which a forward SMPS (that is, an SMPS using a forward converter) is controlled by the TDA2640 to provide the supply voltage for a horizontal deflection circuit. Fig.2 shows a control loop in which a mains-isolated flyback SMPS (that is, an SMPS using a flyback converter) is controlled by a TDA2581 to provide both the supply and drive voltages for a horizontal deflection circuit. In both circuits, a feedback voltage, which is proportional to the SMPS output voltage, is compared with a reference level. The resulting error voltage controls the width of the pulses delivered to the SMPS drive transistor. Since the output voltage of a forward SMPS is δV_i and that from a flyback SMPS with a transformer ratio of n is $\delta V_i/n(1-\delta)$, variation of the output pulse duty factor (δ) changes the voltage transfer of the SMPS to compensate for output load and mains voltage variations.

A low-resistance base drive for the SMPS output transistor is achieved by using transformer coupling between the drive and output transistors. The two transistors should be switched in the non-simultaneous mode for the following reasons:

1) the turn-off conditions, which are critical for high-voltage transistors, are better defined;
2) if the supply for the driver transistor is obtained via a resistor, the base current of the output transistor can be made almost independent of variations in the drive pulse duty factor;
3) the supply for the oscillator can be obtained from the emitter of the driver transistor; (this is not illustrated in Fig.2).

The output pulses from the TDA2640 and the TDA2581 are therefore negative-going for application to the base of the inverting drive transistor TR_1.

In Fig.2, a TDA2571 reconstitutes the received horizontal sync pulses so that they are free from noise and interference, and are independent of sync pulse compression in the transmitted signal. In the TDA2581, these pulses are compared with the flyback pulses from the horizontal deflection circuit. A signal, which is proportional in amplitude to the phase difference, is used to

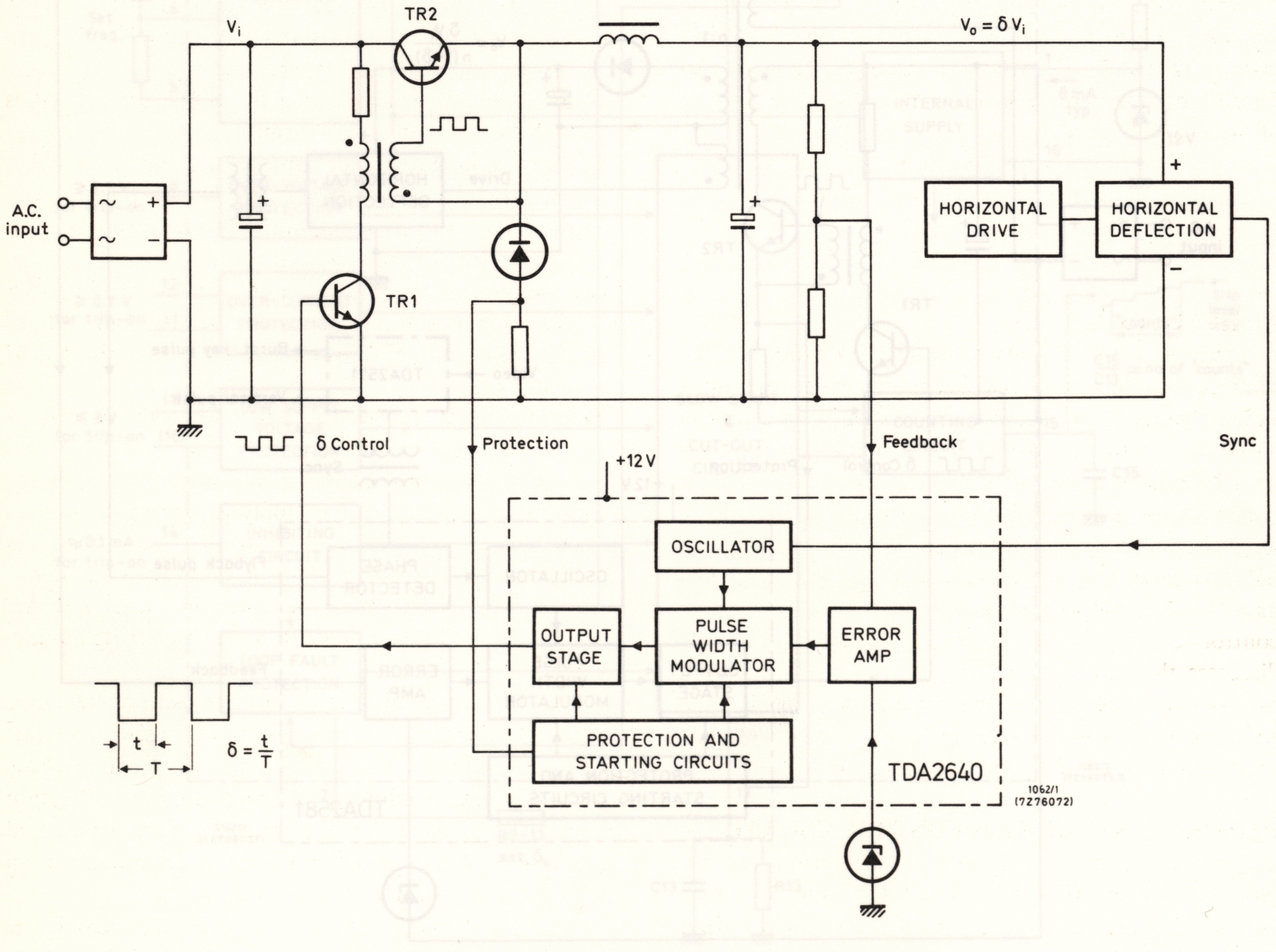


Fig.1 – TDA2640 controlling an SMPS (not mains isolated) using a forward converter to supply power to a horizontal deflection circuit

control the period of a sawtooth waveform from a voltage-controlled line oscillator. The TDA2581 thus produces line-frequency output pulses that are phase-compensated for the variable delays occurring between the line oscillator and the horizontal deflection circuit. When such pulses are used to drive an SMPS to provide the d.c. power for a tv receiver, correctly-timed horizontal deflection drive pulses can also be derived from the SMPS, by adding an additional secondary winding to its output choke.

It should be noted that the circuit arrangements shown in Figs.1 and 2 are only two of the many possible configurations that may be used (Ref.1).

INTEGRATED SMPS CONTROL CIRCUIT TDA2640

The block diagram of the TDA2640 is shown in Fig.3, and the following pin-by-pin functional description provides a basic explanation.

Pin 1: 12V positive supply

The maximum positive supply voltage that may be applied at pin 1 is 13.8V. If the voltage is unstabilised, a voltage regulator diode must be connected between pins 1 and 16 to prevent this maximum from being exceeded. If the voltage on pin 1 falls below about 8V, the low-supply-voltage protection circuit will operate and switch off the output pulses. Current consumption is typically 8mA (maximum 10.5mA).

Pin 2: sync pulse input

Sync pulses applied to pin 2 must be positive-going with an amplitude of between 1 and 10V. The free-running frequency of the integrated oscillator must be higher than the frequency of the sync pulses. The minimum duration of the sync pulses is the difference between the period of the oscillator pulses and the period of the sync pulses. Synchronisation reduces the maximum obtainable duty factor. If synchronisation is not required, link pin 2 to pin 16 (common return).

Pins 3, 4, and 5: oscillator timing network

The timing network consists of a capacitor connected between pins 3 and 4, and a resistor connected between pins 4 and 5. The values of these two components determine the switching period of the SMPS drive pulses. The resistor must be greater than 30kΩ, and curves published in the TDA2640 data sheet enable the correct values for resistor and capacitor to be calculated.

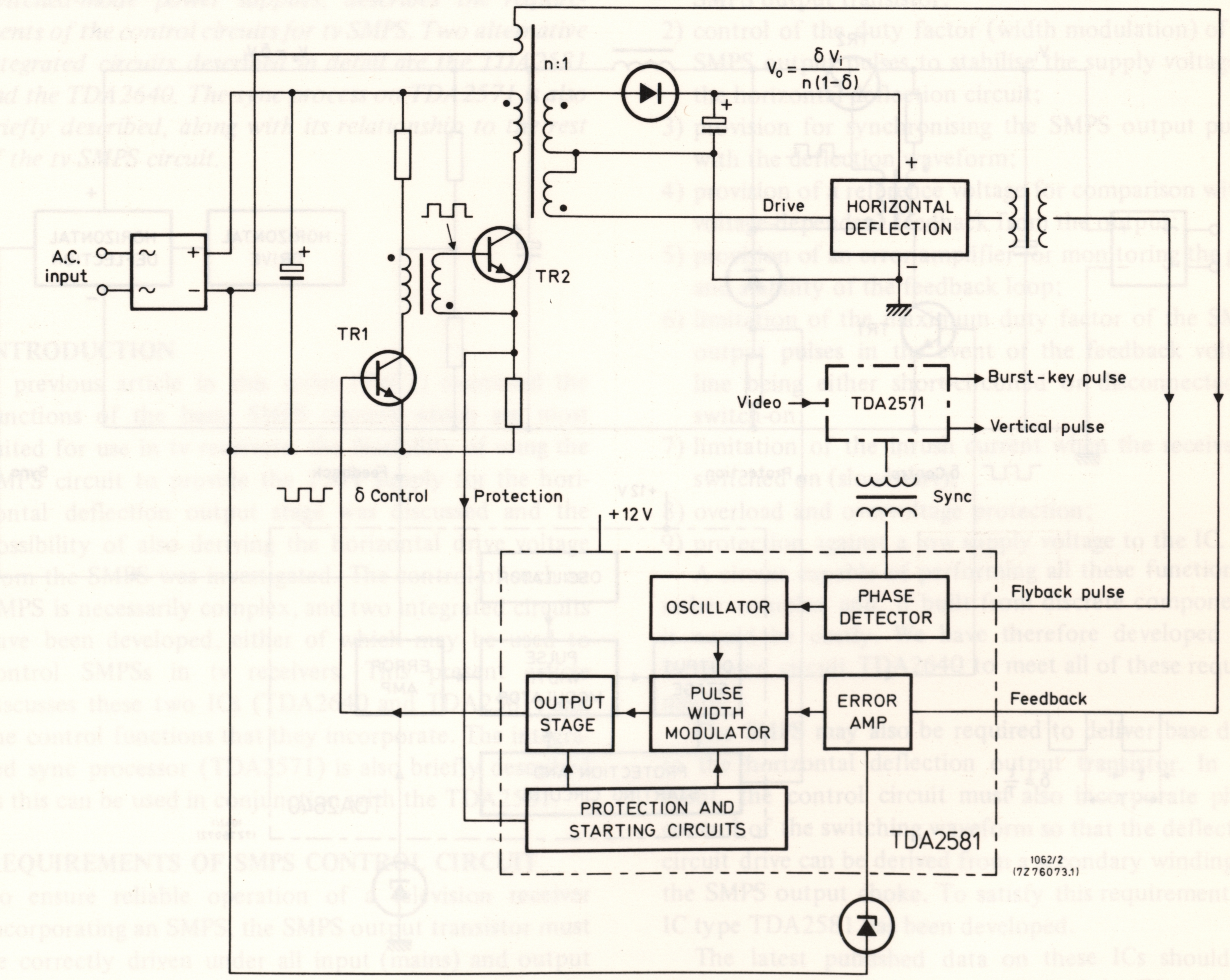

n:1
$V_o = \frac{\delta V_i}{n(1-\delta)}$
A.C. input
TR1
TR2
Drive
HORIZONTAL DEFLECTION
δ Control
Protection
Video
TDA2571
Burst-key pulse
Vertical pulse
Sync
+12V
OSCILLATOR
PHASE DETECTOR
Flyback pulse
OUTPUT STAGE
PULSE WIDTH MODULATOR
ERROR AMP
Feedback
PROTECTION AND STARTING CIRCUITS
TDA2581
1062/2 (7Z76073.1)

Fig.2 – TDA2581 controlling a mains-isolated SMPS using a flyback converter to supply power to a horizontal deflection circuit

Pin 6: open collector output

An external resistor connected between pin 6 and the positive supply rail determines the base drive of the SMPS drive transistor. The integrated output circuit consists of an npn transistor with a catching diode connected between its collector and an internal 12V supply. This provides a low impedance in the ON state (that is, with the drive transistor turned off).

Pin 7: loop fault protection

An external resistor connected between pin 7 and pin 13 determines the maximum obtainable duty factor for the output pulses if the feedback remains below the specified limit during starting.

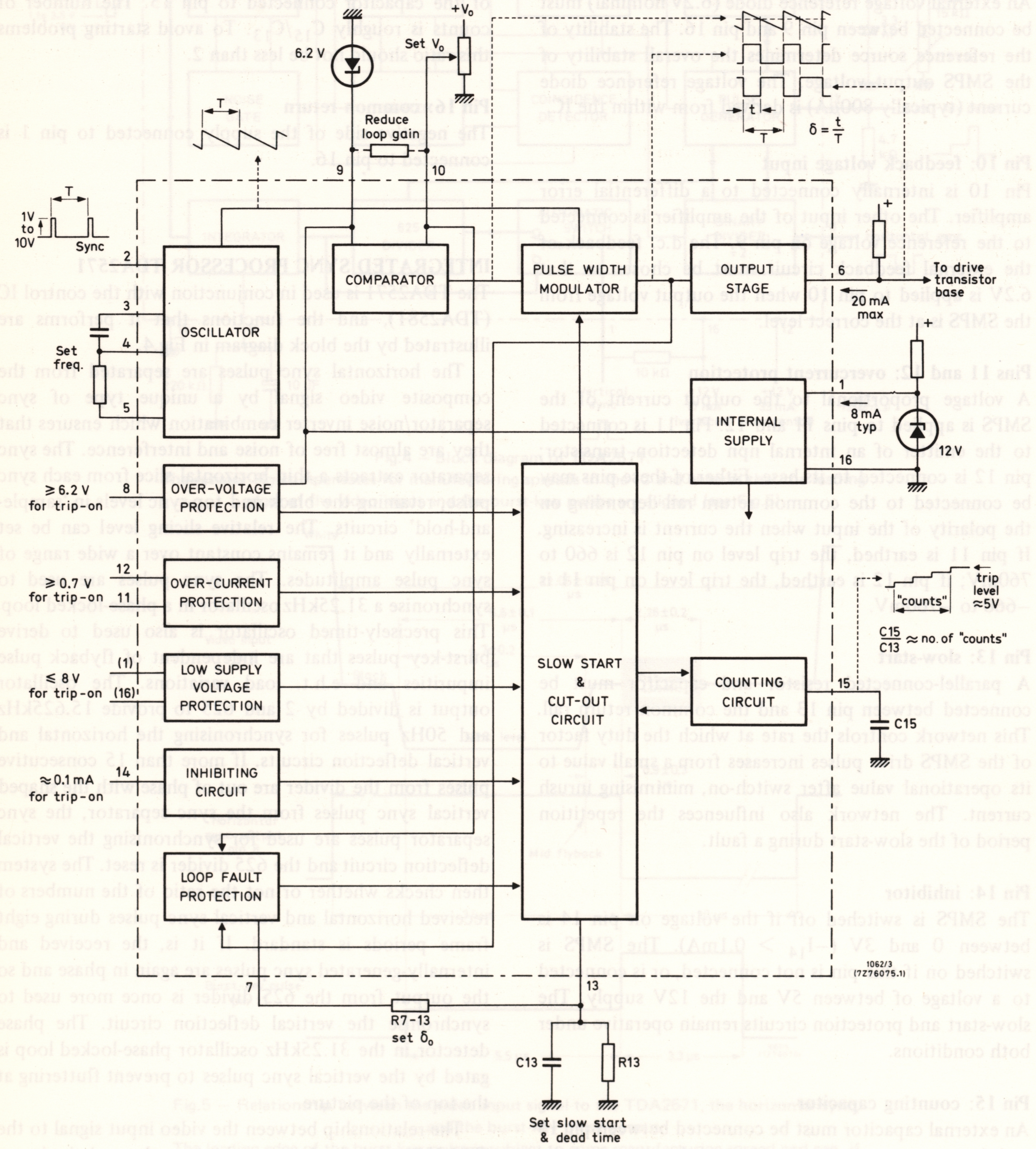

+V₀
Set V₀
6.2V
T
Reduce loop gain
9
10
T
1V to 10V
Sync
2
3
4
5
Set freq.
OSCILLATOR
COMPARATOR
PULSE WIDTH MODULATOR
OUTPUT STAGE
6
To drive transistor base
20mA max
t
T
δ = t/T
+
1
8mA typ
12V
INTERNAL SUPPLY
16
≥6.2V for trip-on
8
OVER-VOLTAGE PROTECTION
12
≥0.7V for trip-on
11
OVER-CURRENT PROTECTION
(1)
≤8V for trip-on
(16)
LOW SUPPLY VOLTAGE PROTECTION
14
≈0.1mA for trip-on
INHIBITING CIRCUIT
LOOP FAULT PROTECTION
SLOW START & CUT-OUT CIRCUIT
COUNTING CIRCUIT
15
C15
trip level ≈5V
"counts"
C15/C13 ≈ no. of "counts"
7
13
R7-13 set δ₀
C13
R13
Set slow start & dead time
1062/3 (7Z76075.1)

Fig.3 – Block diagram of TDA2640

Pin 8: overvoltage protection
A voltage that is proportional to the SMPS output voltage can be connected to pin 8 to operate a protection circuit if a threshold voltage is exceeded. The threshold level is determined by the external voltage reference diode connected to pin 9 (6.2V nominal). If overvoltage protection is not required, pin 8 should be connected to pin 16 (common return).

Pin 9: reference input
An external voltage reference diode (6.2V nominal) must be connected between pin 9 and pin 16. The stability of the reference source determines the overall stability of the SMPS output voltage. The voltage reference diode current (typically 800μA) is derived from within the IC.

Pin 10: feedback voltage input
Pin 10 is internally connected to a differential error amplifier. The other input of the amplifier is connected to the reference voltage on pin 9. The d.c. feedback of the external feedback circuit must be chosen so that 6.2V is applied to pin 10 when the output voltage from the SMPS is at the correct level.

Pins 11 and 12: overcurrent protection
A voltage proportional to the output current of the SMPS is applied to pins 11 and 12. Pin 11 is connected to the emitter of an internal npn detection transistor; pin 12 is connected to its base. Either of these pins may be connected to the common return rail depending on the polarity of the input when the current is increasing. If pin 11 is earthed, the trip level on pin 12 is 660 to 760mV; if pin 12 is earthed, the trip level on pin 11 is −660 to −760mV.

Pin 13: slow-start
A parallel-connected resistor and capacitor must be connected between pin 13 and the common return rail. This network controls the rate at which the duty factor of the SMPS drive pulses increases from a small value to its operational value after switch-on, minimising inrush current. The network also influences the repetition period of the slow-start during a fault.

Pin 14: inhibitor
The SMPS is switched off if the voltage on pin 14 is between 0 and 3V ($-I_{14} > 0.1$mA). The SMPS is switched on if this pin is not connected, or is connected to a voltage of between 5V and the 12V supply. The slow-start and protection circuits remain operative under both conditions.

Pin 15: counting capacitor
An external capacitor must be connected between pin 15 and the common return rail to control the characteristics of the protection circuits.

When the protection circuit operates as a result of the presence of a fault condition, the duty factor of the drive pulses is reduced to zero. After an interval (determined by the time-constant of the circuit connected to pin 13) the duty factor of the pulses slowly increases to its operational value. If the fault is persistent, the duty factor of the pulses is again reduced to zero and the protection sequence is repeated. The number of times that this cycle is repeated before the SMPS drive pulses are permanently discontinued is determined by the value of the capacitor connected to pin 15. The number of counts is roughly C_{15}/C_{13}. To avoid starting problems this ratio should not be less than 2.

Pin 16: common return
The negative side of the supply connected to pin 1 is connected to pin 16.

INTEGRATED SYNC PROCESSOR TDA2571

The TDA2571 is used in conjunction with the control IC (TDA2581), and the functions that it performs are illustrated by the block diagram in Fig.4.

The horizontal sync pulses are separated from the composite video signal by a unique type of sync separator/noise inverter combination which ensures that they are almost free of noise and interference. The sync separator extracts a thin horizontal slice from each sync pulse, retaining the black and top-sync levels in 'sample-and-hold' circuits. The relative slicing level can be set externally and it remains constant over a wide range of sync pulse amplitudes. The sync pulses are used to synchronise a 31.25kHz oscillator in a phase-locked loop. This precisely-timed oscillator is also used to derive burst-key pulses that are independent of flyback pulse impurities and e.h.t. load variations. The oscillator output is divided by 2 and 625 to provide 15.625kHz and 50Hz pulses for synchronising the horizontal and vertical deflection circuits. If more than 15 consecutive pulses from the divider are out of phase with the shaped vertical sync pulses from the sync separator, the sync separator pulses are used for synchronising the vertical deflection circuit and the 625 divider is reset. The system then checks whether or not the ratio of the numbers of received horizontal and vertical sync pulses during eight frame periods is standard. If it is, the received and internally-generated sync pulses are again in phase and so the output from the 625 divider is once more used to synchronise the vertical deflection circuit. The phase detector in the 31.25kHz oscillator phase-locked loop is gated by the vertical sync pulses to prevent fluttering at the top of the picture.

The relationship between the video input signal to the TDA2571 and the horizontal sync pulse and the burst-key pulse outputs is shown in Fig.5.

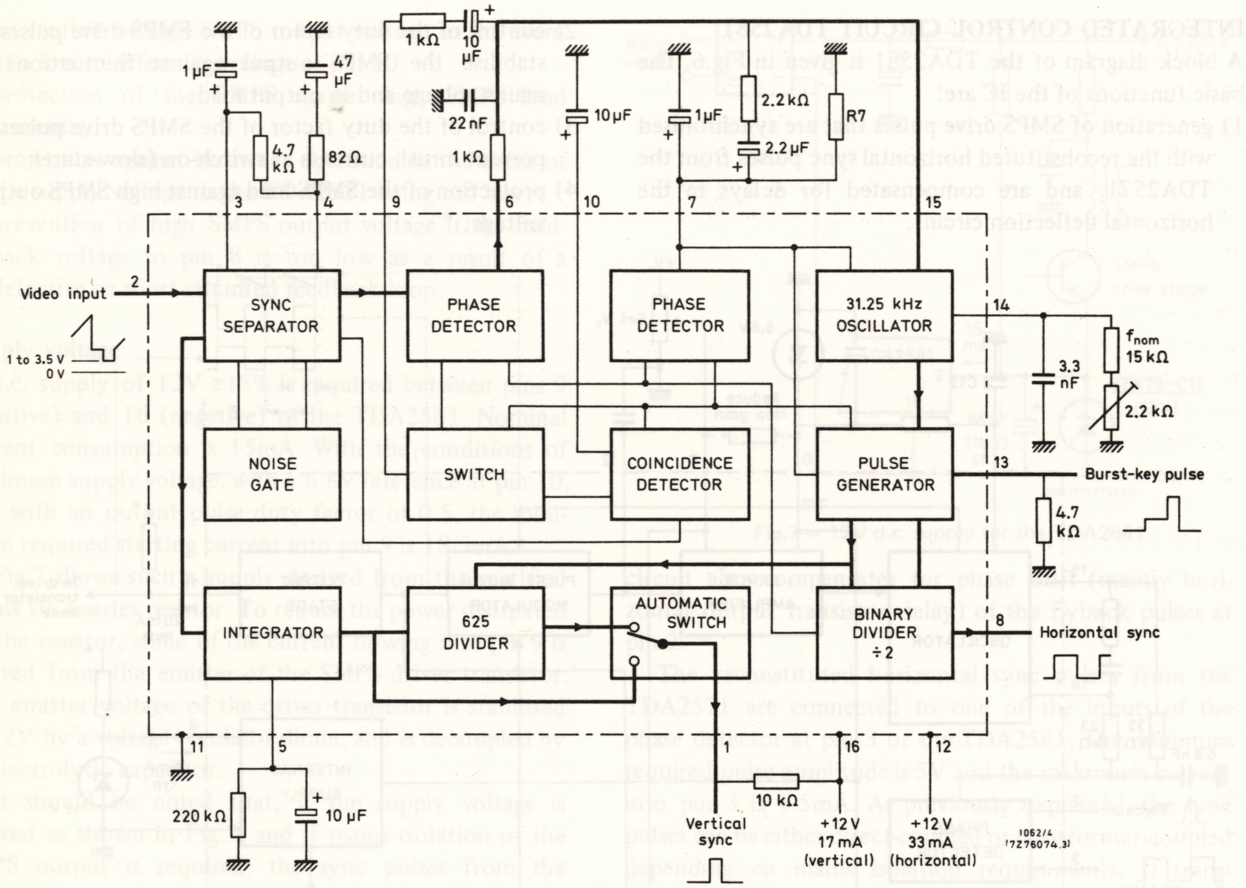


Fig.4 – Block diagram of TDA2571.
The resistor R_7 compensates for manufacturing spreads in the IC and ensures that the leading edges of the video input and the burst-key pulses are aligned (see Fig.5)

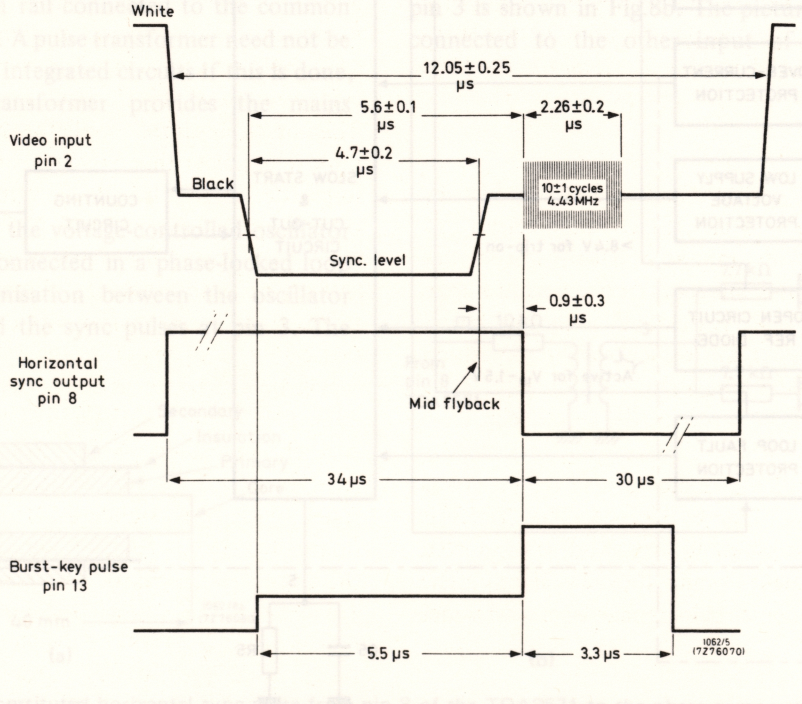
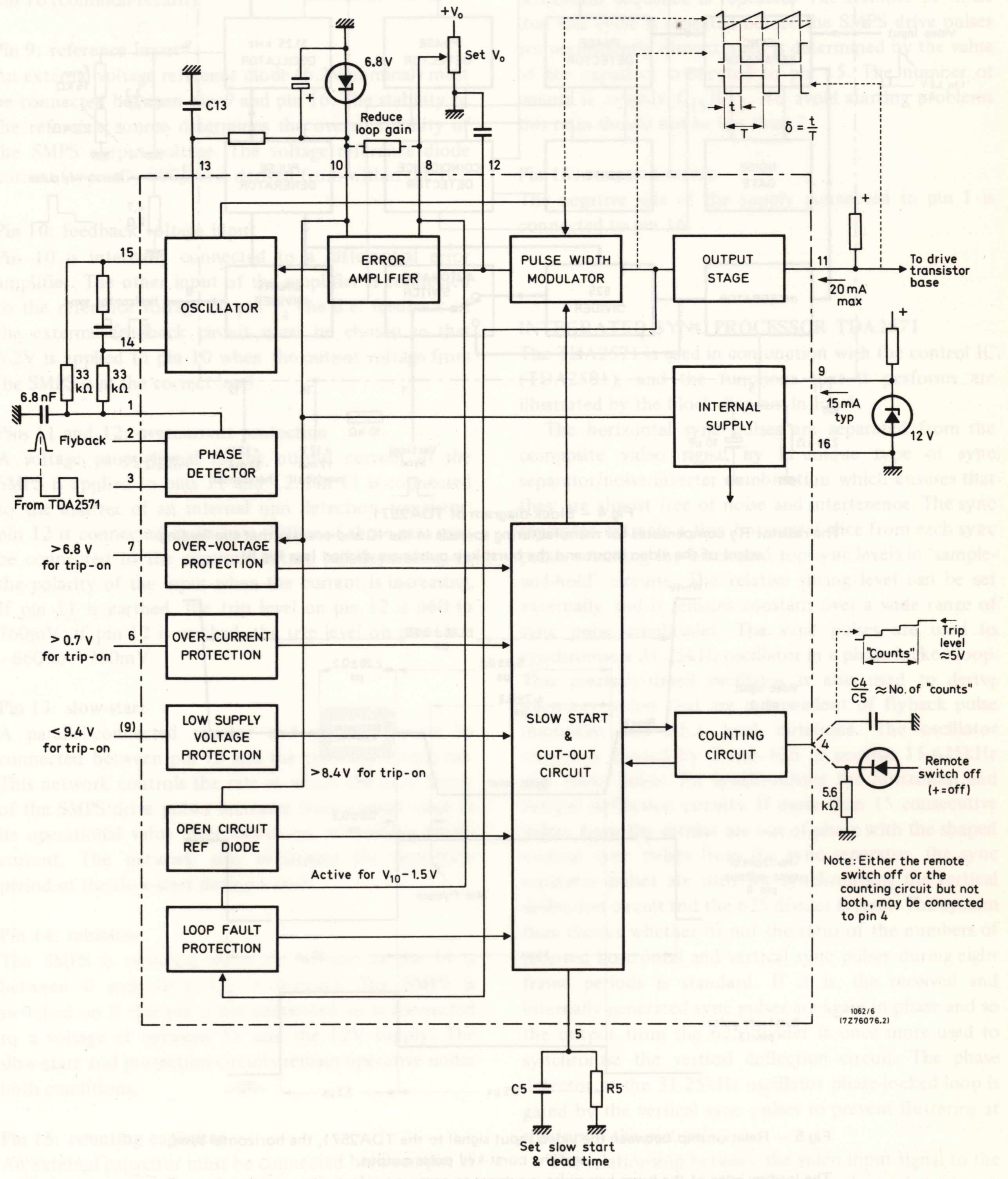


Fig.5 – Relationship between the video input signal to the TDA2571, the horizontal sync, and the burst-key pulse output.
The leading edge of the burst-key pulse is subject to some manufacturing spread and can, if necessary, be aligned with the leading edge of the sync pulse at pin 2 by connecting a resistor of between 68 and 100kΩ between pin 7 and earth (R_7, Fig.4)

INTEGRATED CONTROL CIRCUIT TDA2581

A block diagram of the TDA2581 is given in Fig.6. The basic functions of the IC are:

1) generation of SMPS drive pulses that are synchronised with the reconstituted horizontal sync pulses from the TDA2571, and are compensated for delays in the horizontal deflection circuit;
2) control of the duty factor of the SMPS drive pulses to stabilise the SMPS output against fluctuations in mains voltage and in output load;
3) control of the duty factor of the SMPS drive pulses to prevent inrush currents at switch-on (slow-start);
4) protection of the SMPS load against high SMPS output voltage;



+V₀
Set V₀
6.8 V
C13
Reduce loop gain
13
10
8
12
t
T
δ = t/T
ERROR AMPLIFIER
PULSE WIDTH MODULATOR
OUTPUT STAGE
11
To drive transistor base
20 mA max
OSCILLATOR
15
14
33 kΩ
33 kΩ
6.8 nF
1
2
Flyback
PHASE DETECTOR
3
From TDA2571
INTERNAL SUPPLY
9
15 mA typ
16
12 V
>6.8 V for trip-on
7
OVER-VOLTAGE PROTECTION
>0.7 V for trip-on
6
OVER-CURRENT PROTECTION
"Counts"
Trip level ≈5V
C4/C5 ≈ No. of "counts"
<9.4 V for trip-on
(9)
LOW SUPPLY VOLTAGE PROTECTION
>8.4V for trip-on
SLOW START & CUT-OUT CIRCUIT
COUNTING CIRCUIT
4
Remote switch off (+=off)
5.6 kΩ
OPEN CIRCUIT REF DIODE
Active for V₁₀–1.5 V
LOOP FAULT PROTECTION
1062/6 (7Z76076.2)
5
C5
R5
Set slow start & dead time

Note: Either the remote switch off or the counting circuit but not both, may be connected to pin 4

Fig.6 – Block diagram of TDA2581

5) prevention of high SMPS output voltage in the event of a loss of reference voltage at pin 10;
6) protection of the SMPS against high SMPS load currents;
7) protection against incorrect operation in the event of the supply voltage to the TDA2581 being too low;
8) prevention of high SMPS output voltage if the feedback voltage to pin 8 is too low as a result of a defective or short-circuited feedback loop.

Supply voltage

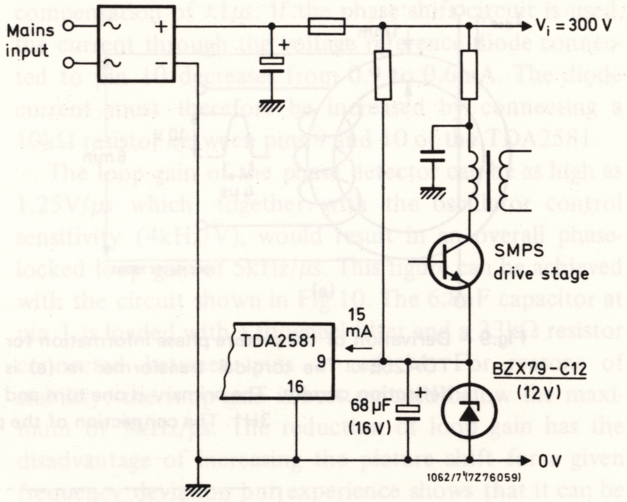
A d.c. supply of 12V ±15% is required between pins 9 (positive) and 16 (negative) of the TDA2581. Nominal current consumption is 15mA. With the conditions of minimum supply voltage, with a 6.8V reference at pin 10, and with an output pulse duty factor of 0.5, the minimum required starting current into pin 9 is 18.5mA.

Fig.7 shows such a supply derived from the rectified mains via a series resistor. To reduce the power dissipated by the resistor, some of the current flowing into pin 9 is derived from the emitter of the SMPS driver transistor. The emitter voltage of the driver transistor is stabilised at 12V by a voltage regulator diode, and is decoupled by an electrolytic capacitor.

It should be noted that, if the supply voltage is derived as shown in Fig.7, and if mains isolation of the SMPS output is required, the sync pulses from the TDA2571 must be connected to the input of the TDA2581 via a small pulse transformer. The 12V supply can however be derived from a separate mains transformer and its common rail connected to the common rail of the SMPS output. A pulse transformer need not be connected between the integrated circuits if this is done, because the driver transformer provides the mains isolation.

Phase detector

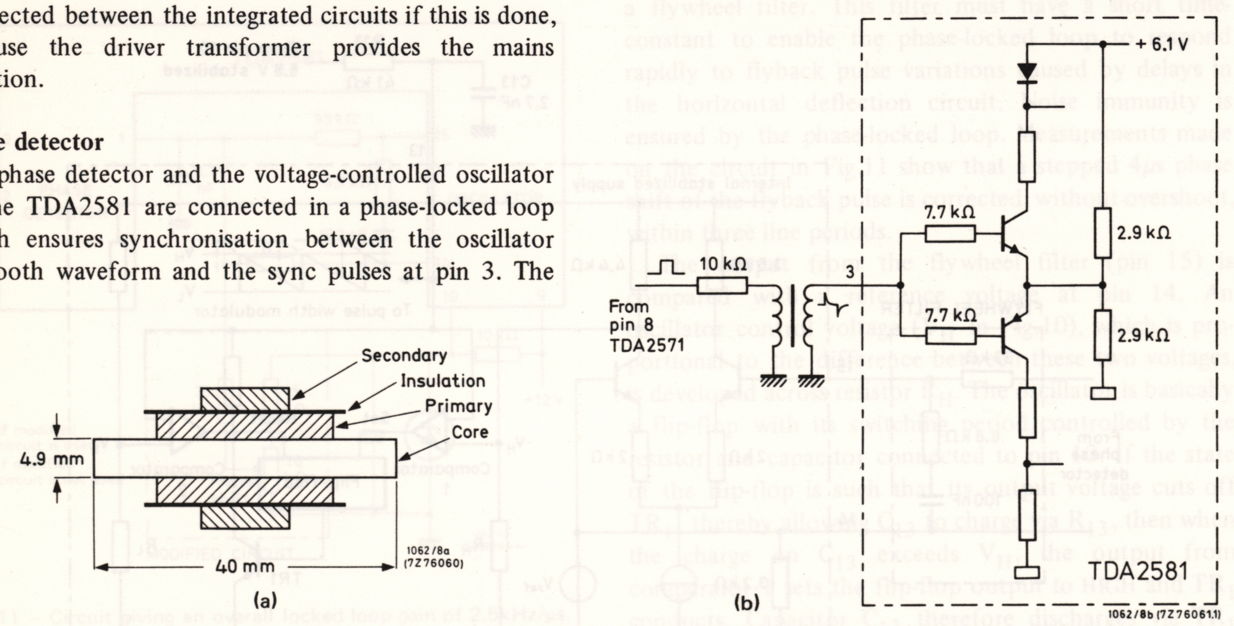
The phase detector and the voltage-controlled oscillator of the TDA2581 are connected in a phase-locked loop which ensures synchronisation between the oscillator sawtooth waveform and the sync pulses at pin 3. The circuit also compensates for phase shift (mainly horizontal output transistor delay) of the flyback pulses at pin 2.



Mains input
Vᵢ = 300 V
SMPS drive stage
15 mA
TDA2581
9
16
BZX79-C12 (12 V)
68 µF (16 V)
0 V
1062/7(7Z76059)

Fig.7 – 12V d.c. supply for the TDA2581

The reconstituted horizontal sync pulses from the TDA2571 are connected to one of the inputs of the phase detector at pin 3 of the TDA2581. The minimum required pulse amplitude is 5V and the maximum current into pin 3 is 1.5mA. As previously explained, the sync pulses can be either direct-coupled or transformer-coupled depending on mains isolation requirements. If transformer coupling is required, a suitable transformer can be constructed as shown in Fig.8a. The part of the internal circuit of the TDA2581 that is connected to pin 3 is shown in Fig.8b. The picture phase information connected to the other input of the phase detector



Secondary
Insulation
Primary
Core
4.9 mm
40 mm
1062/8a (7Z76060)
(a)
+ 6.1 V
7.7 kΩ
2.9 kΩ
10 kΩ
3
From pin 8 TDA2571
7.7 kΩ
2.9 kΩ
TDA2581
(b)
1062/8b (7Z76061.1)

Fig.8 – Connecting the reconstituted horizontal sync pulse from pin 8 of the TDA2571 to the phase comparator input (pin 3) of the TDA2581 when mains isolation is required. A cross-section of the transformer is shown in (a). The core is a rod of Ferroxcube 3C6 (Mullard A9). The primary is 65 turns of 0.25mm diameter wire wound in one layer and has an inductance of 145µH. The secondary is 25 turns of the same wire, again in one layer. The transformer connections are as shown in (b)

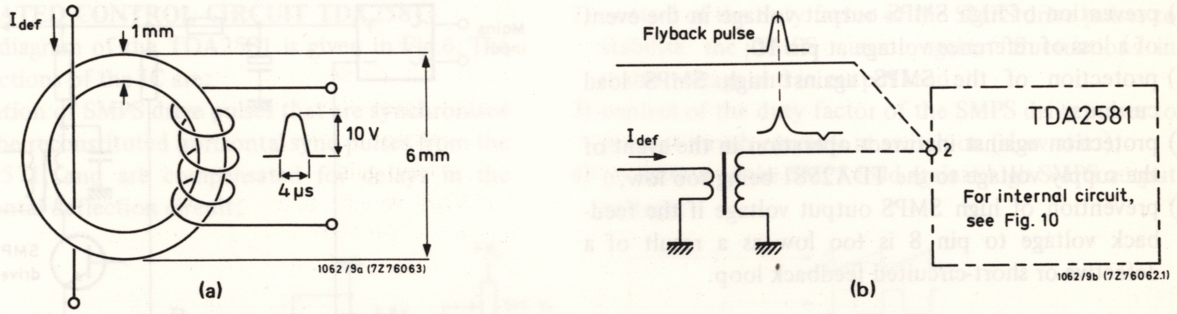

I_{def}
1mm
10 V
4 μs
6mm
1062/9a (7Z76063)
(a)
Flyback pulse
I_{def}
TDA2581
2
For internal circuit, see Fig. 10
1062/9b (7Z76062.1)
(b)

Fig.9 – Derivation of the picture phase information for application to one of the inputs of the phase detector in the TDA2581. The toroidal transformer in (a) is for obtaining a pulse representing the mid-scan from the deflection current. The primary is one turn and the secondary is 15 turns, wound on a toroid of Ferroxcube 3H1. The connection of the picture phase information is shown in (b)

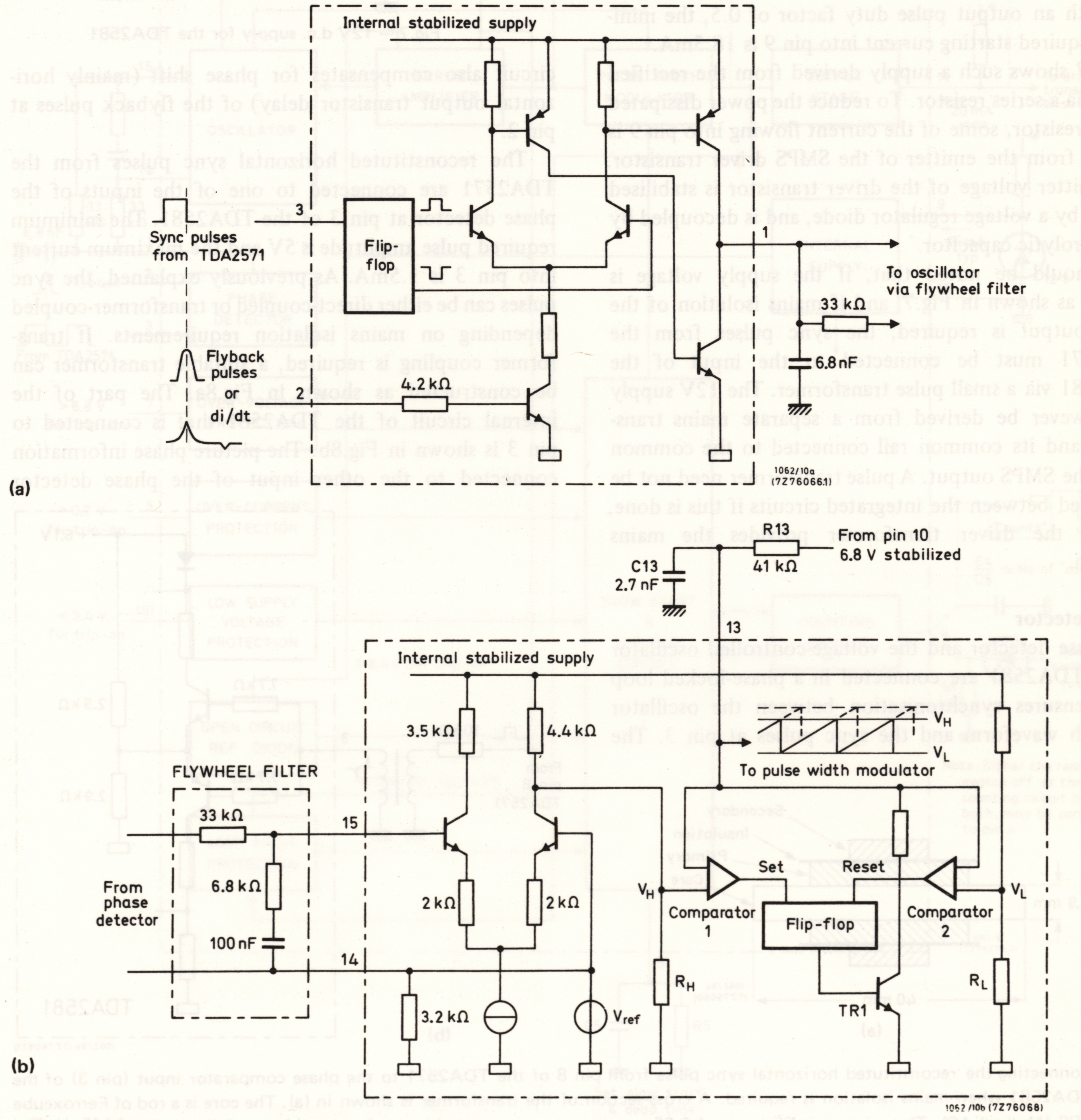

Internal stabilized supply
3
Sync pulses from TDA2571
Flip-flop
1
To oscillator via flywheel filter
33 kΩ
6.8 nF
Flyback pulses or di/dt
4.2 kΩ
2
(a)
1062/10a (7Z76066)
R13
41 kΩ
From pin 10 6.8 V stabilized
C13 2.7 nF
13
Internal stabilized supply
3.5 kΩ
4.4 kΩ
V_H
V_L
To pulse width modulator
FLYWHEEL FILTER
33 kΩ
15
From phase detector
6.8 kΩ
100 nF
14
2 kΩ
2 kΩ
3.2 kΩ
V_{ref}
Set
Reset
V_H
V_L
Comparator 1
Flip-flop
Comparator 2
R_H
R_L
TR1
(b)
1062/10b (7Z76068)

Fig.10 – TDA2581: (a) phase detector;
(b) oscillator

(pin 2) must be a pulse with a duration of at least 3μs to maintain the control sensitivity. The maximum permitted current into pin 2 is 1.5mA. The input to pin 2 can be a normal flyback pulse, or a spike obtained from a small toroidal core that is saturated at all times except when the horizontal deflection current passes through zero (Fig.9).

A simplified diagram of the phase detector is given in Fig.10a. The sync pulses from the TDA2571 trigger a flip-flop squaring circuit connected to pin 3 of the TDA2581. The flip-flop allows differentiated pulses to be accepted at pin 3, thereby permitting the use of a simple pulse transformer if mains isolation is required. The phase detector compares the output from the flip-flop with the picture phase information from pin 2, and switches a bidirectional current source in accordance with any phase differential. The current source output swing (+1.0 to −1.5mA) controls the charge on the capacitor connected to pin 1.

The current source is switched at the negative-going edges of the sync pulses applied to pin 3, and the output capacitor is charged to an average level of 1.5V. The delay between the middle of the flyback information and the negative edge of the sync pulse is 0.9μs ± 0.3μs (see Fig.5). The unequal positive and negative currents from the phase detector output compensate for this delay. A small change in the value of the output capacitor may be necessary, however, to compensate for the delay in the small-signal stages of the television receiver (0.3μs maximum) and eccentricity of the electron beam in the picture tube. The circuit shown in Fig.11 adjusts the charge level to provide a phase compensation of ±1μs. If the phase shift circuit is used, the current through the voltage reference diode connected to pin 10 decreases from 0.9 to 0.6mA. The diode current must therefore be increased by connecting a 10kΩ resistor between pins 9 and 10 of the TDA2581.

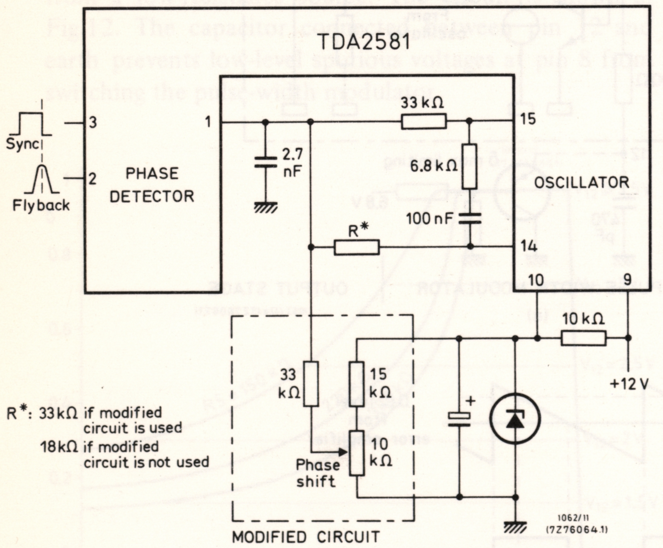

TDA2581
Sync
3
Flyback
2
PHASE DETECTOR
1
2.7 nF
33 kΩ
15
6.8 kΩ
OSCILLATOR
100 nF
R*
14
10
9
10 kΩ
+12 V
33 kΩ
15 kΩ
10 kΩ
Phase shift
MODIFIED CIRCUIT
R*: 33kΩ if modified circuit is used
18kΩ if modified circuit is not used
1062/11 (7Z76064.1)

Fig.11 – Circuit giving an overall locked loop gain of 2.5kHz/μs. The modification shown introduces a phase shift between the sync and the flyback pulses at the input of the TDA2581 phase detector. Note that the 18kΩ resistor has to be increased to 33kΩ if the modification circuit is incorporated

The loop gain of the phase detector can be as high as 1.25V/μs which, together with the oscillator control sensitivity (4kHz/V), would result in an overall phase-locked loop gain of 5kHz/μs. This figure can be achieved with the circuit shown in Fig.10. The 6.8nF capacitor at pin 1 is loaded with a flywheel filter and a 33kΩ resistor connected between pins 1 and 14. For reasons of stability, the loop gain is best reduced below the maximum of 5kHz/μs. The reduction of loop gain has the disadvantage of increasing the picture shift for a given frequency deviation but experience shows that it can be reduced to a value of 1kHz/μs without an unacceptable degradation in performance.

The circuit shown in Fig.11 gives an optimum compromise between stability and picture shift. The overall phase-locked loop gain is 2.5kHz/μs with or without the modification shown: the modification shown in Fig.11 enables an introduced phase shift to be adjusted (by means of the 10kΩ potentiometer) so that the sync and flyback pulses at the input of the TDA2581 phase detector may be synchronised.

Flywheel filter and oscillator

A simplified circuit diagram of the oscillator is given in Fig.10b.

The voltage across the capacitor at the output from the phase detector (pin 1) is applied to the oscillator via a flywheel filter. This filter must have a short time-constant to enable the phase-locked loop to respond rapidly to flyback pulse variations caused by delays in the horizontal deflection circuit. Noise immunity is ensured by the phase-locked loop. Measurements made on the circuit in Fig.11 show that a stepped 4μs phase shift of the flyback pulse is corrected, without overshoot, within three line periods.

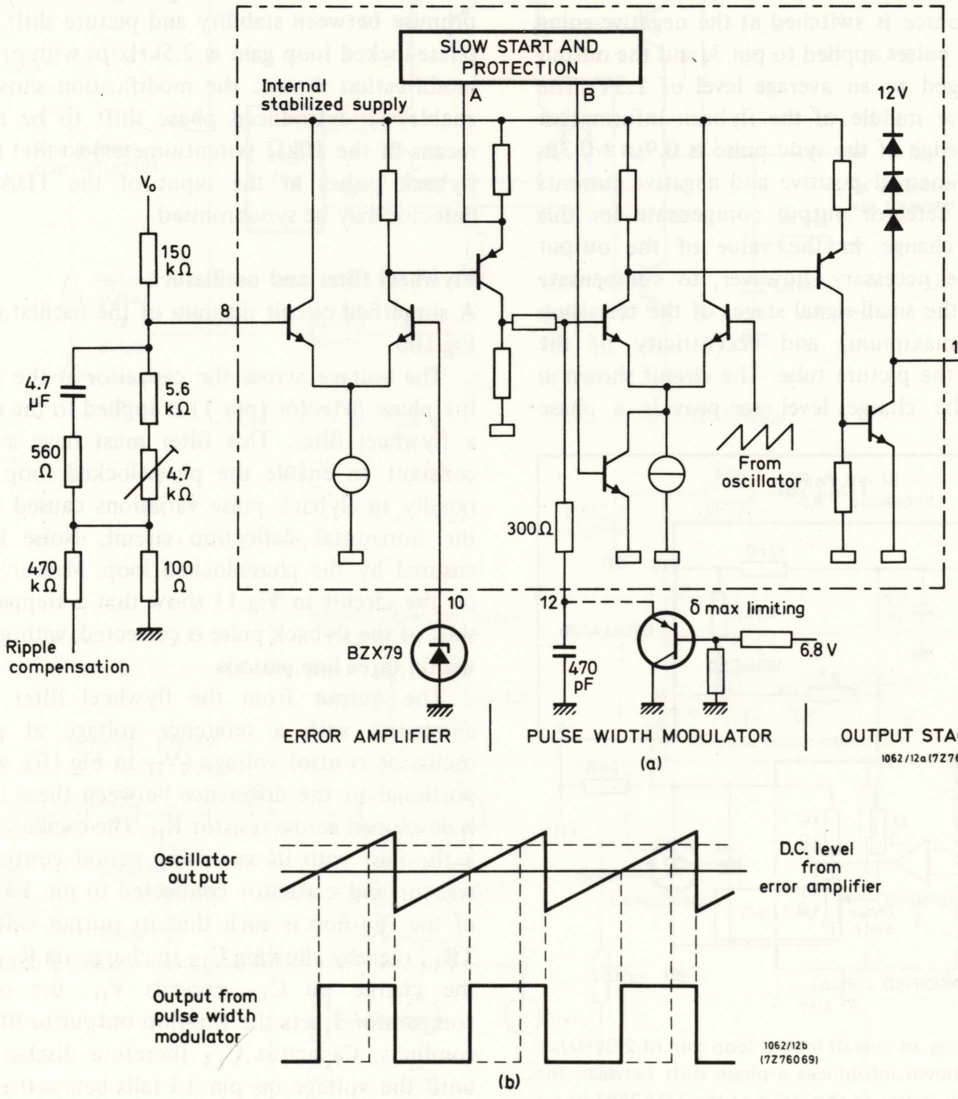
The output from the flywheel filter (pin 15) is compared with a reference voltage at pin 14. An oscillator control voltage (V_H in Fig.10), which is proportional to the difference between these two voltages, is developed across resistor R_H. The oscillator is basically a flip-flop with its switching period controlled by the resistor and capacitor connected to pin 13. If the state of the flip-flop is such that its output voltage cuts off TR_1, thereby allowing C_{13} to charge via R_{13}, then when the charge on C_{13} exceeds V_H, the output from comparator 1 sets the flip-flop output to HIGH and TR_1 conducts. Capacitor C_{13} therefore discharges via TR_1 until the voltage on pin 13 falls below the voltage V_L. The output from comparator 2 then sets the flip-flop output to LOW, thereby cutting off TR_1 and allowing

C_{13} to recharge via R_{13}. The oscillator frequency can be varied ±1.5kHz from nominal by adjusting the control voltage V_H.

The output from pin 13 is a sawtooth waveform with an amplitude excursion from 1.45 to 3.95V. With R_{13} = 41kΩ (a preset potentiometer is suggested) and C_{13} = 1.7nF, the nominal period of the oscillator is 64μs. If 1% tolerance components are used for R_{13} and C_{13}, and the voltage reference diode has a 5% tolerance, the maximum frequency spread is limited to ±3%; frequency adjustment is then not necessary. The charging current for C_{13} is derived from the voltage regulator diode connected to pin 10; the maximum frequency deviation for a reference voltage change from 6 to 7V is −2%. The nominal frequency cannot be checked by disconnecting the pulses from pins 2 and 3 of the TDA2581 because the output from the phase detector would then be undefined. The conversion gain of the oscillator is 4.5kHz/V and the catching range is ±1.5kHz. The typical temperature stability is −300ppm/°C.

Error amplifier

A simplified circuit diagram of the error amplifier is given in Fig.12. The circuit amplifies the difference between the fraction of the SMPS output voltage at pin 8 and the voltage across the voltage regulator diode (5.6 to 7.5V) connected to pin 10. The current available from pin 10 for a 6.8V diode is 0.9mA. If the diode needs a higher current to operate in the stabilising mode, a resistor can be connected between pin 10 and the 12V supply rail. Mains ripple compensation can be incorporated by applying a ripple signal to pin 8 as shown. The output from the error amplifier is a d.c. level proportional to the deviation of the SMPS output voltage away from its set level. The output is fed, via a buffer stage, to the pulse-width modulator. The loop gain can be reduced (see the Appendix) by connecting a resistor between the inputs to the error amplifier (pins 8 and 10).



(a)

(b)

Fig.12 – Simplified circuit of error amplifier, pulse-width modulator, and output amplifier with associated waveforms

Pulse-width modulator

A simplified circuit diagram of the pulse-width modulator is also given in Fig.12. The circuit produces an output when the amplitude of the oscillator sawtooth at one input is less than the output from the error amplifier applied to the other input. The waveforms in Fig.12 show that the output pulses from the circuit have a width that is proportional to the output level from the error amplifier. The duty factor of the pulses is therefore proportional to the amount by which the SMPS output voltage (V_o) deviates from its set value.

Output stage

The final part of the circuit shown in Fig.12 is the output stage which is driven by a preamplifier. The collector of the output transistor is protected against the application of high voltages to pin 11 by three internal clamping diodes. The amplitude of the pulses across the external collector load resistor at pin 11 is dependent on the clamping level and on the knee voltage of the output transistor. These pulses have a leading-edge rise time of 0.1μs and a duty factor of up to 0.98. The output voltage will not exceed 0.4V if the output current remains below 20mA.

Fig.13 shows the duty factor of the pulses from the pulse-width modulator as a function of SMPS output voltage sample (V_8). The diagram also shows that the maximum value for the duty factor can be limited by connecting pin 12 via a clamping diode to a d.c. level from a low-resistance source. The circuit is shown in Fig.12. The capacitor connected between pin 12 and earth prevents low-level spurious voltages at pin 8 from switching the pulse-width modulator.

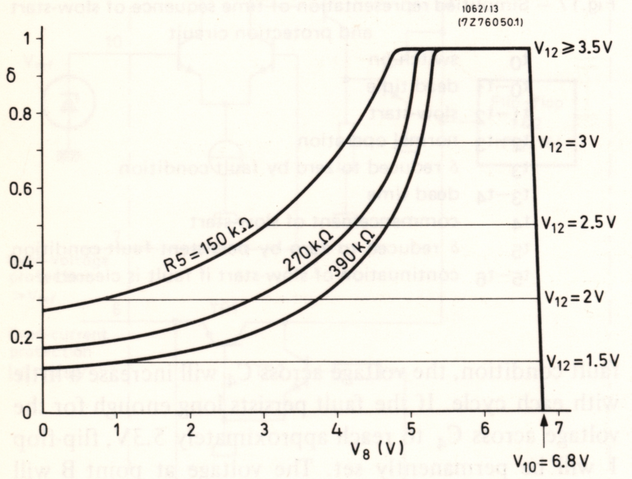


Fig.13 – Duty factor of the TDA2581 output pulses as a function of V_8 with R_5 as a parameter and with different limiting values of V_{12}

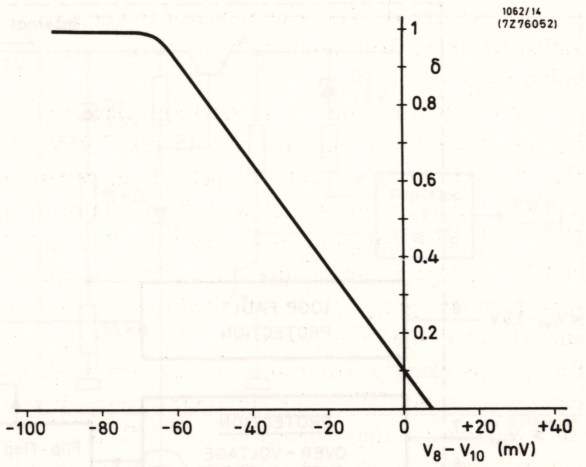


Fig.14 – Duty factor of the TDA2581 output pulses as a function of the error amplifier input. This curve represents the near-vertical lines at the right of Fig.13 on an expanded horizontal scale

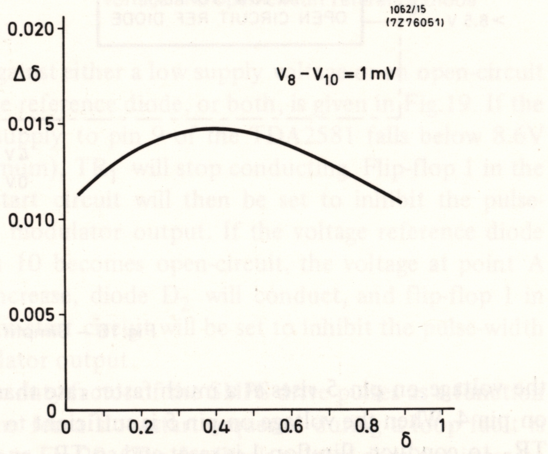


Fig.15 – Change of duty factor of the TDA2581 output pulses for a 1mV error amplifier input change, as a function of initial duty factor. This curve shows the non-linearity of the nearly-straight diagonal line in Fig.14

Fig.14 and 15 clarify the operation of the circuit when the difference between the reference voltage and the output sample voltage is very small.

Slow-start and protection circuit

To prevent inrush current from damaging the output transistor of the SMPS when the receiver is switched on, it is necessary to ensure that the duty factor of the SMPS drive pulses increases slowly from zero to the normal operation value: this is known as slow start.

A simplified diagram of the slow-start and protection circuit is given in Fig.16, and a timing diagram is given in Fig.17.

When the receiver is switched on, flip-flop 1 applies a LOW signal to the bases of TR_2 and TR_4 and so they both conduct. A HIGH level from point B of the same flip-flop, applied to the pulse-width modulator, inhibits the SMPS drive pulses. Capacitors C_4 and C_5 charge, and

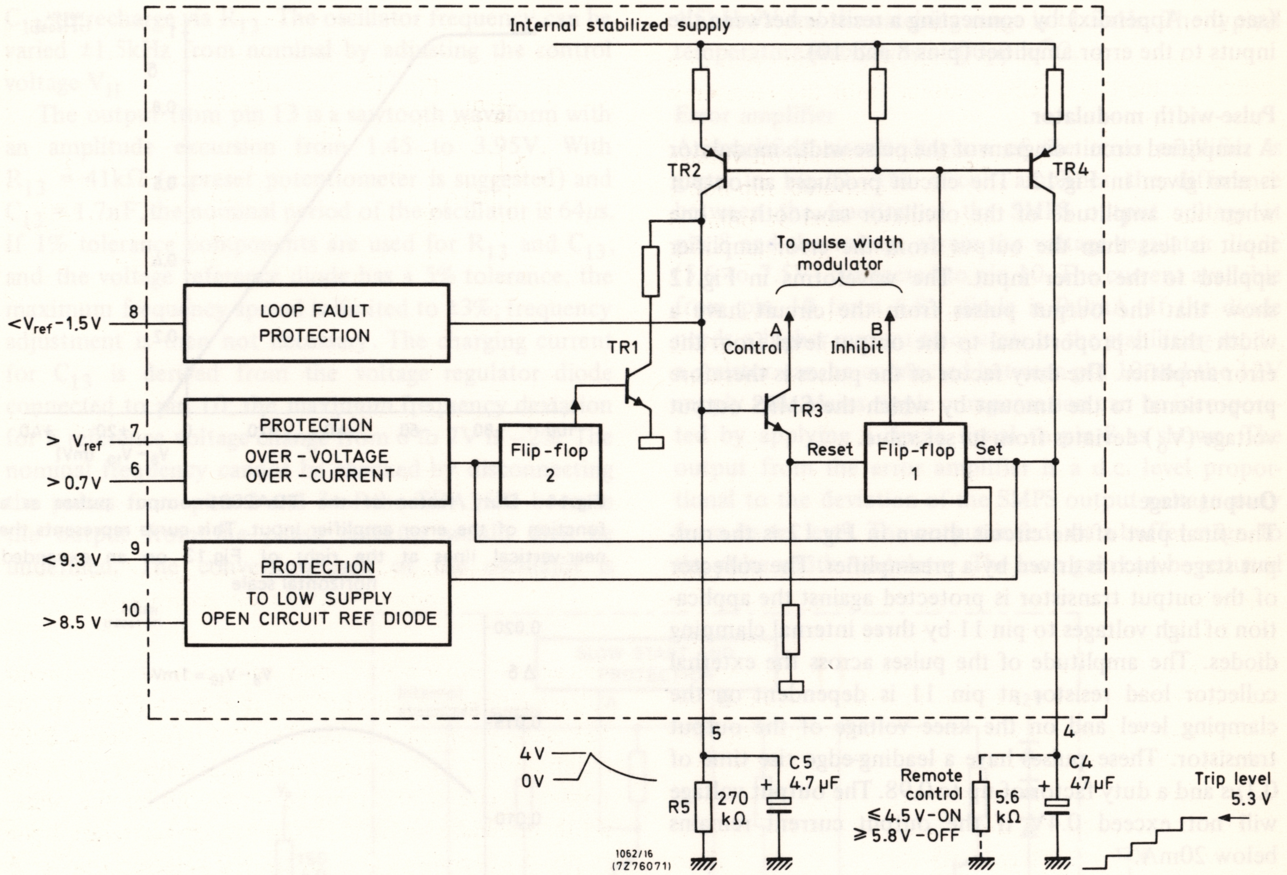


Fig.16 — Simplified slow-start and protection circuit

the voltage on pin 5 rises at a much faster rate than that on pin 4. When the voltage on pin 5 is sufficient to cause TR_3 to conduct, flip-flop 1 is reset and so TR_2 and TR_4 cease to conduct. The LOW signal at point A ensures that the pulse-width modulator output pulses remain inhibited. As C_5 discharges via R_5, the slowly increasing voltage at point A allows the duty factor of the pulse-width modulator output pulses to increase slowly to the operational value dictated by the output from the error amplifier. The rate of duty factor increase (slow-start period) is dictated by the values of R_5 and C_5.

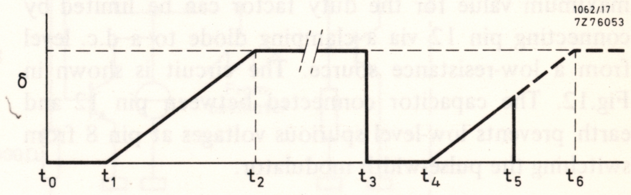


Fig.17 — Simplified representation of time sequence of slow-start and protection circuit

t_0	switch-on
t_0–t_1	dead time
t_1–t_2	slow-start
t_2–t_3	normal operation
t_3	δ reduced to zero by fault condition
t_3–t_4	dead time
t_4	commencement of slow-start
t_5	δ reduced to zero by persistent fault condition
t_5–t_6	continuation of slow-start if fault is cleared

If the overvoltage or overcurrent protection circuit is activated, flip-flop 1 is again set to the starting condition and the output from the pulse-width modulator is again inhibited. Simultaneously flip-flop 2 is set and its output causes TR_1 to conduct and to clamp the voltage across C_5 to a low value. When the fault is corrected (either because the fault was transient or because the SMPS output reaches zero), flip-flops 1 and 2 are reset and TR_1 ceases to conduct, thus enabling C_5 to charge and initiate the slow-start sequence once again. If the fault persists, the protection circuit will again activate during the slow-start sequence and the cycle will be repeated. Each time flip-flop 1 is set, C_4 charges briefly through TR_4. During a sequence of slow-start cycles caused by a persistent fault condition, the voltage across C_4 will increase a little with each cycle. If the fault persists long enough for the voltage across C_4 to reach approximately 5.3V, flip-flop 1 will be permanently set. The voltage at point B will then ensure that the output pulses from the pulse-width modulator remain inhibited until the receiver is switched off and then on again. If the 'counting' action is not required, C_4 should be replaced by a 5.6kΩ resistor; the retriggering of the slow-start sequence in the event of a persistent fault will then continue until the receiver is switched off.

If the low supply voltage or the open-circuit voltage regulator diode protection circuit is activated, flip-flop 1 is permanently set and the voltage level at point B ensures that the output pulses from the pulse-width modulator remain inhibited.

An overload condition at the SMPS output or a fault in the feedback path can produce a voltage at pin 8 which is too low. A protection circuit will, in this event, reduce the voltage at pin 5 to limit the duty factor of the SMPS drive pulses. The relationship between the pulse duty factor and the voltage on pin 8 is shown in Fig.13.

Protection input circuits

The overvoltage or overcurrent protection input circuits are shown in simplified form in Fig.18. The voltage at pin 7 is compared with the reference voltage, and any excess is amplified and used for setting both flip-flops in the slow-start circuit (Fig.16). Flip-flop 1 inhibits the output pulses from the pulse-width modulator. Flip-flop 1 limits the maximum duty factor of the SMPS drive pulses when the slow-start sequence recommences.

The flip-flops in the slow-start circuit are also set if the voltage at pin 6 exceeds the V_{BE} of TR_1 (−660 to −760mV) or of TR_2 (660 to 760mV). The overcurrent protection can thus be initiated by either a positive or a negative voltage. The current levels for operation are −100 or 2µA respectively.

A simplified version of the circuit which offers protection against either a low supply voltage or an open-circuit voltage reference diode, or both, is given in Fig.19. If the 12V supply to pin 9 of the TDA2581 falls below 8.6V (minimum), TR_1 will stop conducting. Flip-flop 1 in the slow-start circuit will then be set to inhibit the pulse-width modulator output. If the voltage reference diode at pin 10 becomes open-circuit, the voltage at point A will increase, diode D_2 will conduct, and flip-flop 1 in the low-start circuit will be set to inhibit the pulse-width modulator output.

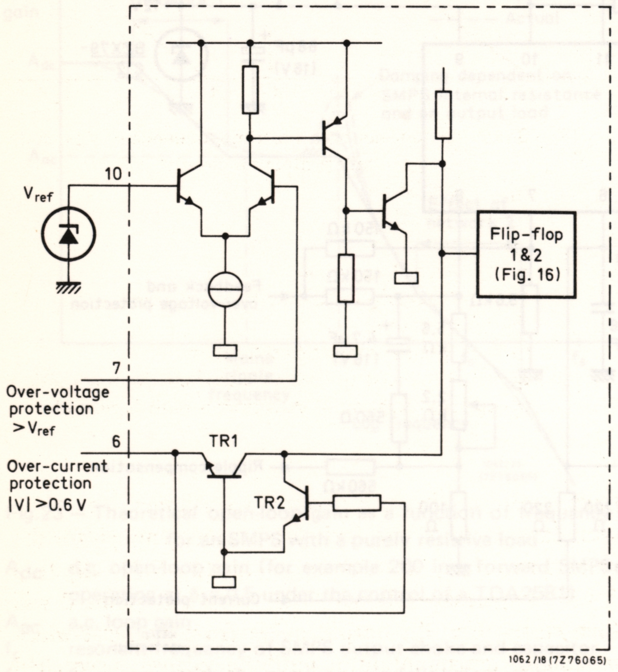

V$_{ref}$
10
7
Over-voltage protection >V$_{ref}$
6
Over-current protection |V|>0.6 V
TR1
TR2
Flip-flop 1&2 (Fig. 16)
1062/18 (7Z76065)

Fig.18 — Overvoltage and overcurrent protection

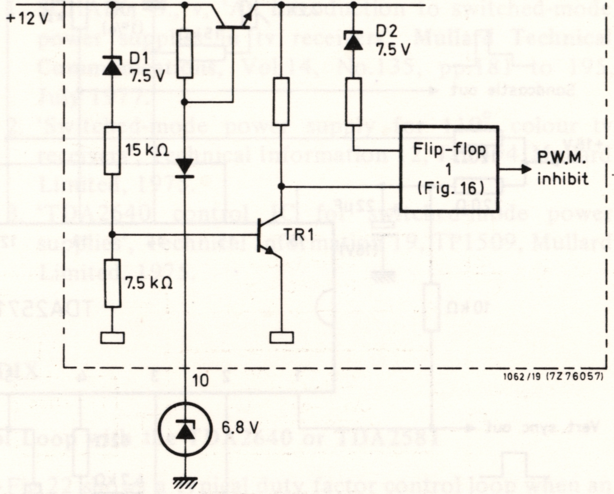


Fig.19 — Simplified circuit for protection against low supply voltage and open-circuit reference diode

The duty factor of the SMPS drive pulses as a function of time for a slow-start sequence during a loop fault is shown in Fig.20. (The feedback voltage does not increase if pin 8 of the TDA2581 is disconnected or short-circuited to earth). The slow-start sequence will commence in the normal manner but will be arrested when the duty factor of the drive pulses reaches a small value. If the loop fault

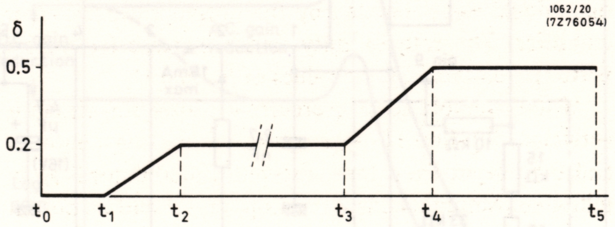


Fig.20 — Protection against loop fault (loss of feedback) during starting

t_0	switch-on
t_0–t_1	dead time
t_1–t_2	commencement of slow-start
t_2–t_3	δ limited to a low value due to a loop fault
t_3	loop fault corrected
t_3–t_4	continuation of slow-start
t_4	δ for normal operation reached
t_4-t_5	normal operation, regulation of power supply taken over by control loop

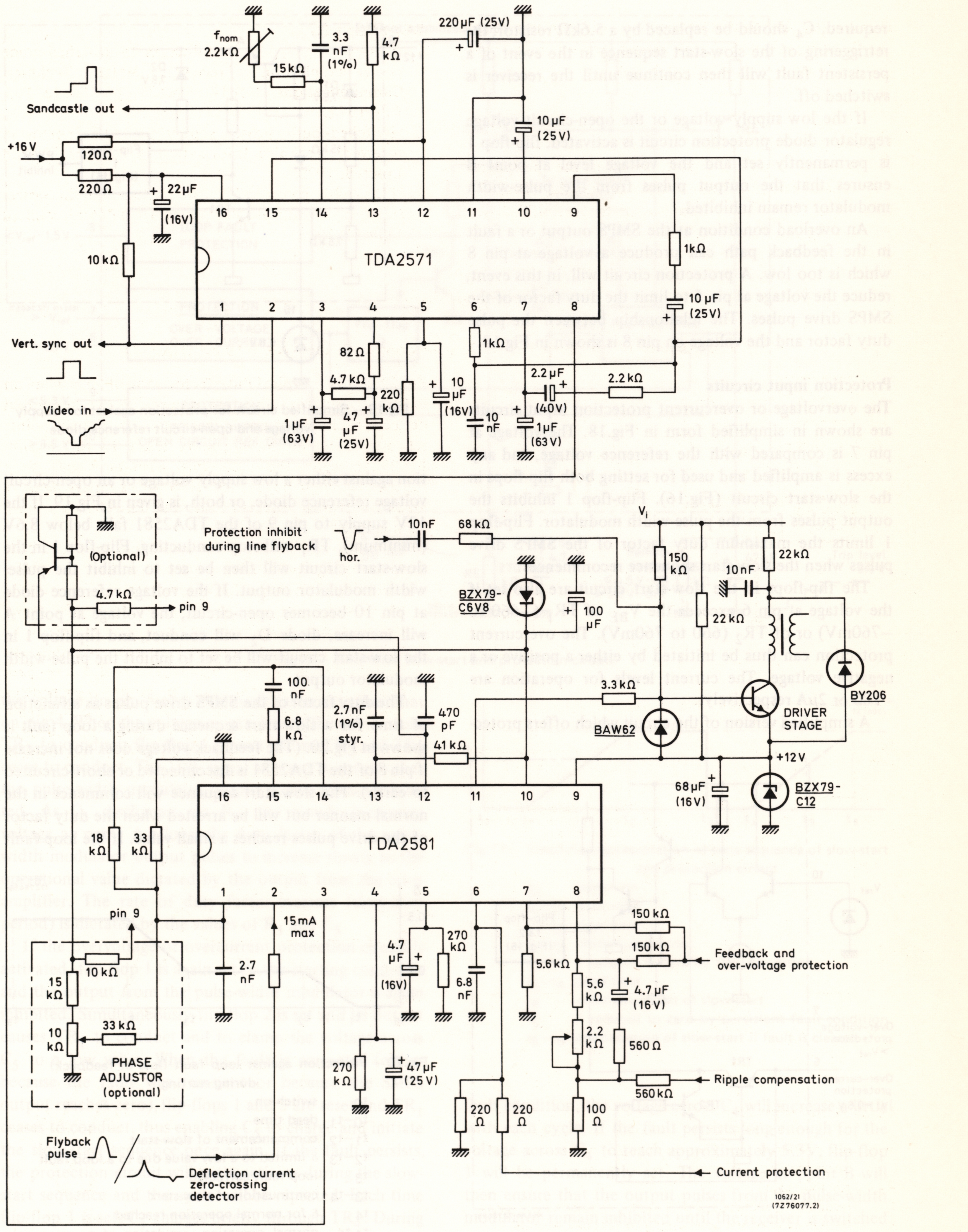

TDA2571
TDA2581
Sandcastle out
+16V
Vert. sync out
Video in
Protection inhibit during line flyback
(Optional)
hor.
pin 9
PHASE ADJUSTOR (optional)
Flyback pulse
Deflection current zero-crossing detector
15mA max
Feedback and over-voltage protection
Ripple compensation
Current protection
DRIVER STAGE
+12V
BZX79-C6V8
BAW62
BY206
BZX79-C12
f_{nom} 2.2kΩ
1062/21 (7Z76077.2)

Fig.21 – TDA2571 and TDA2581 controlling an SMPS driver stage

is corrected, the slow-start sequence will continue until the duty factor of the drive pulses reaches the normal operation value of approximately 0.5.

Complete control circuit

A complete control circuit showing the peripheral components and interconnections between the TDA2571 and TDA2581 is shown connected to an SMPS driver stage in Fig.21.

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2. 'Switched-mode power supply for 110° colour tv receivers', Technical Information 12, TP1494, Mullard Limited, 1975.
3. 'TDA2640 control IC for switched-mode power supplies', Technical Information 19, TP1509, Mullard Limited, 1975.

APPENDIX

Gain and Stability of the Duty Factor Control Loop with the TDA2640 or TDA2581

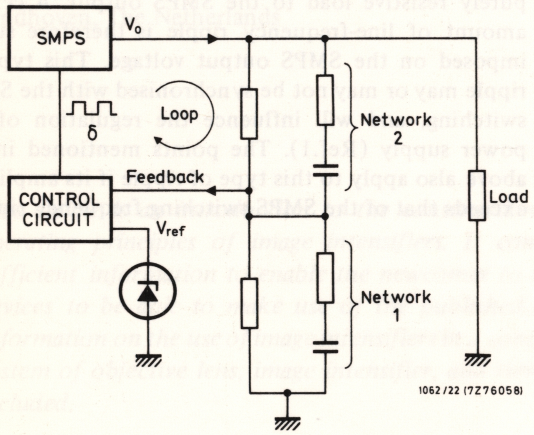
SMPS
V_o
δ
Loop
Feedback
CONTROL CIRCUIT
V_{ref}
Network 2
Network 1
Load
1062/22 (7Z76058)

Fig.22 – Typical SMPS δ control feedback loop

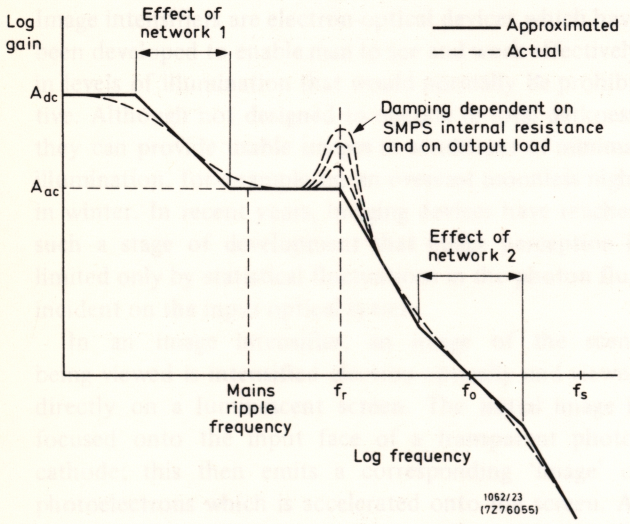

Fig.23 – Theoretical open-loop gain as a function of frequency for an SMPS with a purely resistive load

A_{dc} d.c. open-loop gain (for example 200 in a forward SMPS operating at $\delta = 0.5$ under the control of a TDA2581)

A_{ac} a.c. loop gain

f_r resonant frequency of SMPS output choke and capacitor

f_o frequency at which open-loop gain falls below unity

f_s SMPS switching frequency

Fig.22 shows a typical duty factor control loop when an SMPS is used for providing power for the horizontal deflection circuit of a television receiver: the six points set out below in this appendix relate to this circuit. Fig.23 indicates the theoretical open-loop gain as a function of frequency for an SMPS with a purely resistive load. The storage effects of the switching transistor have been ignored.

1) In theory the frequency at which the loop gain falls to unity f_o must not be more than half the SMPS switching frequency f_s. In practice, f_o in a stable SMPS will be between $0.1f_s$ and $0.2f_s$. Fig.24 shows that if a very high d.c. open-loop gain results in f_o being too high, the required loop-gain reduction can be achieved in one of two ways. The d.c. loop gain can be reduced by, for example, connecting a resistor between the error amplifier inputs. The a.c. loop gain can be reduced by connecting a resistor and capacitor in series between the feedback input to the control circuit and earth (network 1, Fig.22).

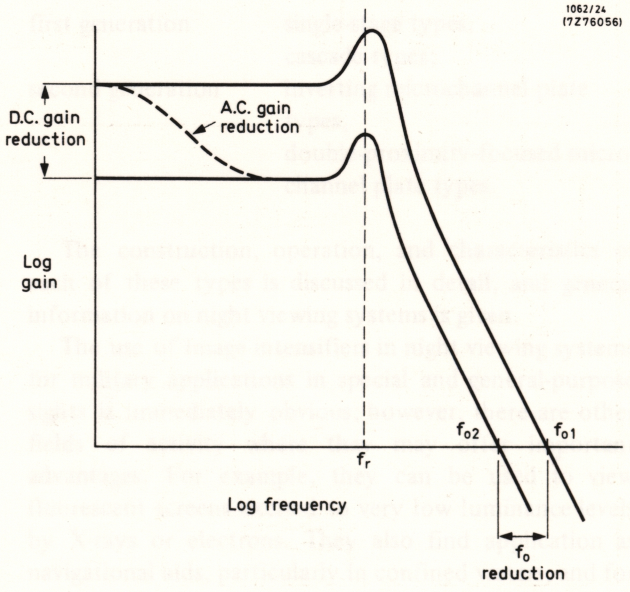

Fig.24 – Reducing f_o by reducing the a.c. or d.c. open-loop gain

2) The loop gain roll-off at high frequencies is caused mainly by the effect of the SMPS output choke and capacitor (LC filter) which causes a phase lag of almost 180°. To ensure stability, the phase lag at f_o must not be more than 180° and must preferably be limited to 135°. The phase lag can be reduced by the use of network 2 in Fig.22.
3) The transient response of the control loop depends upon two factors, as follows.

   Firstly, if the phase lag at f_o is too close to 180°, then high-frequency overshoot and ringing will occur. The phase lag can be reduced as described in (2) above.

   Secondly, if network 1 is used for gain reduction, as in (1) above, it will introduce a low-frequency overshoot. This may lead to excessive voltages in the circuit, for example, during feedback take-over following the slow start sequence.
4) The stability of the SMPS output voltage with respect to slow variations of the rectified mains voltage or to 100Hz ripple on the d.c. input is about equal to the open-loop gain at the variation frequency. If the previously mentioned requirements result in a loop gain that gives insufficient ripple reduction, feedforward (ripple compensation) must also be used.
5) For the pulse-width modulator sensitivity to remain constant, the output from the error amplifier (and therefore the feedback voltage at the error amplifier input) must be smooth. If too much ripple is superimposed on the feedback voltage, instability will occur. This is the most difficult problem to solve when designing a control loop because there will always be a certain amount of switching frequency ripple superimposed on the SMPS output voltage, and any attempt to smooth it will cause phase lag. Although the use of network 1 (a.c. gain reduction) will reduce this type of instability, the use of network 2 (phase lead) tends to increase it. Ideally the loop should incorporate a lowpass filter with a sharp cut-off between f_o and f_s. In any event, the SMPS output capacitor must have a low equivalent series resistance and inductance.
6) A horizontal deflection circuit does not present a purely resistive load to the SMPS output. A certain amount of line-frequency ripple is therefore superimposed on the SMPS output voltage. This type of ripple may or may not be synchronised with the SMPS switching, and will influence the regulation of the power supply (Ref.1). The points mentioned in (5) above also apply to this type of ripple if its amplitude exceeds that of the SMPS switching frequency ripple.

Image intensifiers — an introduction

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This article is an introduction to the construction and operating principles of image intensifiers. It contains sufficient information to enable the newcomer to these devices to be able to make use of the published data. Information on the use of image intensifiers in a complete system of objective lens, image intensifier, and viewer is included.

INTRODUCTION

Image intensifiers are electron-optical devices which have been developed to enable man to see and work effectively in levels of illumination that would normally be prohibitive. Although not designed to operate in total darkness, they can provide usable images in conditions of minimal illumination, for example on an overcast moonless night in winter. In recent years, imaging devices have reached such a stage of development that image perception is limited only by statistical fluctuations in the photon flux incident on the input optical system.

In an image intensifier, an image of the scene being viewed is intensified electron-optically and viewed directly on a luminescent screen. The initial image is focused onto the input face of a transparent photocathode; this then emits a corresponding 'image' of photoelectrons which is accelerated onto the screen. As each photon from the scene is made to produce several tens of photons at the screen, the final image is greatly intensified.

There are two main categories of intensifier.

Active image converters

These are 'active' in that they normally require the scene to be artificially illuminated, for example, by infrared radiation. The device then converts reflected IR to an intensified directly-viewable image.

Passive image intensifiers

These intensify visible and near-infrared radiation. They are 'passive' in that they operate in available light and require no additional illumination. They enable clear vision in conditions of near total darkness.

Many different devices, both electrostatically-focused and magnetically-focused, belonging to one of these two broad categories have been developed. However, in this article, only electrostatically-focused passive image intensifiers are considered. Of these, there are four main types, generally grouped in two generations:

first generation	single-stage types, cascade types;
second generation	inverting microchannel plate types, double-proximity-focused microchannel plate types.

The construction, operation, and characteristics of each of these types is discussed in detail, and general information on night viewing systems is given.

The use of image intensifiers in night viewing systems for military applications in special and general-purpose sights is immediately obvious; however, there are other fields of activity where they may offer important advantages. For example, they can be used to view fluorescent screens excited to very low luminance levels by X-rays or electrons. They also find application as navigational aids, particularly in confined waters, and for surveillance and traffic control, in low ambient light conditions. They also improve the effectiveness of astronomical telescopes. Where it is advantageous for the observer to be remote from the viewing instrument, the intensifier can be coupled optically to a television camera. Such a system allows monitoring in a hazardous environment, or allows a number of observers to view a scene simultaneously, or one observer to monitor several screens.

AIDING THE PERCEPTION OF THE EYE

The average number of photons emitted per second by an element in a night scene is linearly proportional to the brightness of the element. The fraction of these photons that falls on the retina of the eye of an observer depends upon the size of the pupil, that is upon the state of dark adaptation, and upon the distance of the observer from the object. Also, the number of photons received per second from any one element will vary with time as a result of the random emission process.

The process of seeing 'detail' in a scene depends on the ability of the eye to detect a difference in illuminance at adjoining areas on the retina. This can be enhanced by an instrument which:

1) captures a larger fraction of the photons emitted by the scene;
2) has a higher quantum efficiency than that of the unaided eye;
3) has a spectral response which is matched more closely than the eye to the night sky illumination (for example, the S25 photocathode used in our image intensifiers).

The image intensifiers discussed in this publication are intended for use in passive night viewing systems. They rely for their operation on available light at extremely low levels of intensity such as that from the stars or reflected town glow. The table shows the range of scene illuminance (brightness) normally experienced.

Range of Scene Illuminance

Direct sunlight	10^5 lx
Bright daylight	10^4
Overcast day	10^3
Very dark day	10^2
Twilight	10
Deep twilight	1
Full moon	10^{-1}
Quarter moon	10^{-2}
Starlight	10^{-3}
Overcast starlight	10^{-4}

With the aid of good objective lenses, the amount of light reaching the photocathode of an intensifier when the scene is illuminated by overcast starlight will be approximately 10^{-6} lx. An intensifier must be able to sense this very small amount of radiation and intensify it to a level detectable by the eye. Intensifiers must also be usable under twilight conditions when the photocathode illumination is about 0.1 lx. To cope with this range of light levels, a ratio of about 10^5, image intensifiers generally have automatic brightness control (a.b.c.), sometimes called automatic gain control (a.g.c.).

Small very-bright sources of light can create problems. However, image intensifiers using microchannel plates (second-generation types, see later) are able to accept small intense sources and still present a useful image.

FIRST-GENERATION IMAGE INTENSIFIERS

As mentioned earlier, image intensifiers can be divided into two groups, termed first-generation types and second-generation types. First-generation types are described below.

Electrostatically-focused single-stage image intensifiers

An electrostatically-focused single-stage image intensifier is shown schematically in Fig.1. Light from the object is focused onto the input to the tube by means of an objective lens and transmitted through the fibre-optic window onto the semitransparent multi-alkali photocathode. A corresponding electron pattern is emitted. These electrons are then accelerated by a voltage of up to 16kV onto an aluminised phosphor screen. An inverted intensified photon image is generated, and this is transmitted through the output window to appear as a directly-viewable image on the screen of the intensifier. The level of intensification of the image depends mainly on the photocathode sensitivity, the accelerating voltage, and the tube magnification. The typical luminance gain of an intensifier of this type would be between 100 and 2000, depending on the parameters mentioned above. A gain of 2000 is sufficient to give an extremely bright image of a moonlit scene.

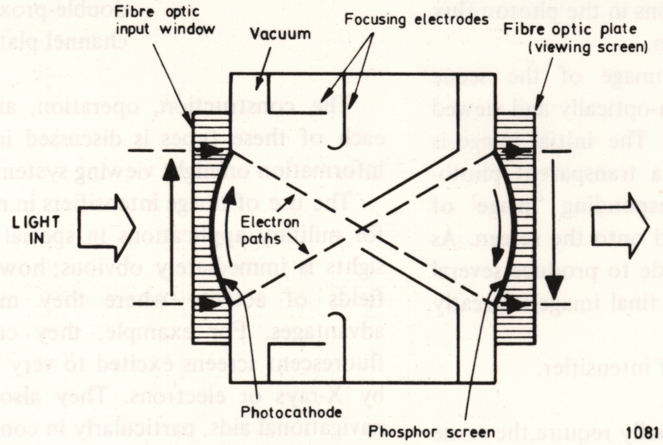


Fig.1 — Electrostatically-focused single-stage image intensifier

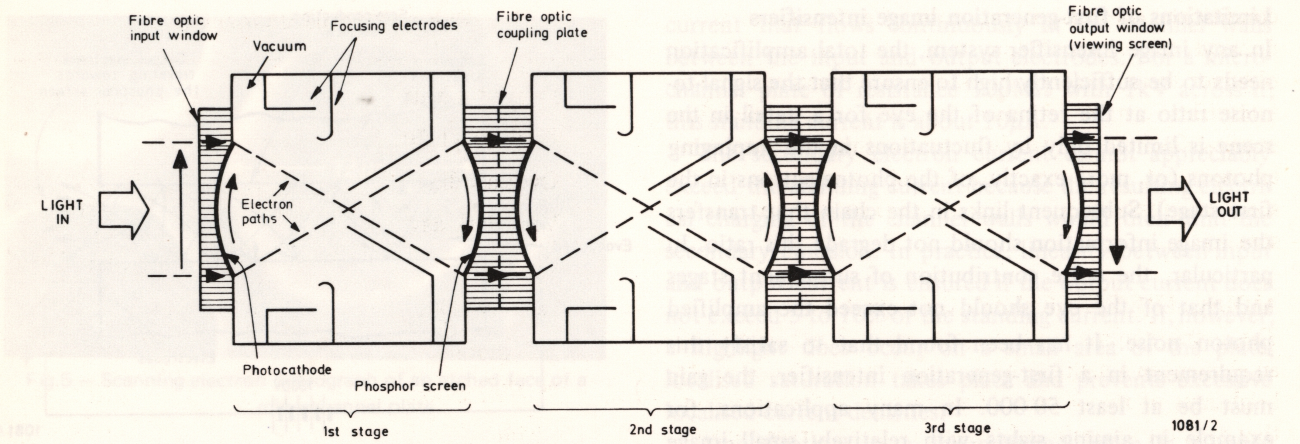


Fig.2 — Electrostatically-focused cascade image intensifier

For scenes illuminated by overcast starlight, theoretical studies show that a luminance gain of 5000 to 10 000 is sufficient to give a useful image with first-generation devices, provided that the observer is fully dark-adapted and is using an optimised system. In fact, extensive field trials and laboratory measurements indicate that for first-generation devices the detail contrast perceptibility (that is, the usefulness of the image) increases with gain up to about 60 000. Beyond 65 000, the increased gain does not contribute to increased information.

Electrostatically-focused cascade image intensifiers

To obtain a gain of 50 000 or greater, single-stage electrostatically-focused intensifiers can be coupled in cascade (see Fig.2). This has been made possible by the development of high-resolution vacuum-tight fibre-optic windows. Typical gains of greater than 50 000 can be achieved with such a configuration.

The three stages of the cascade intensifier together require an e.h.t. of about 45kV. This is derived from an oscillator, encapsulated with the intensifier, operating from a low-voltage d.c. supply such as alkaline batteries. The integral-oscillator cascade image intensifier has a well-defined automatic brightness control characteristic (see Fig.3) and a fast response to sudden changes in scene brightness.

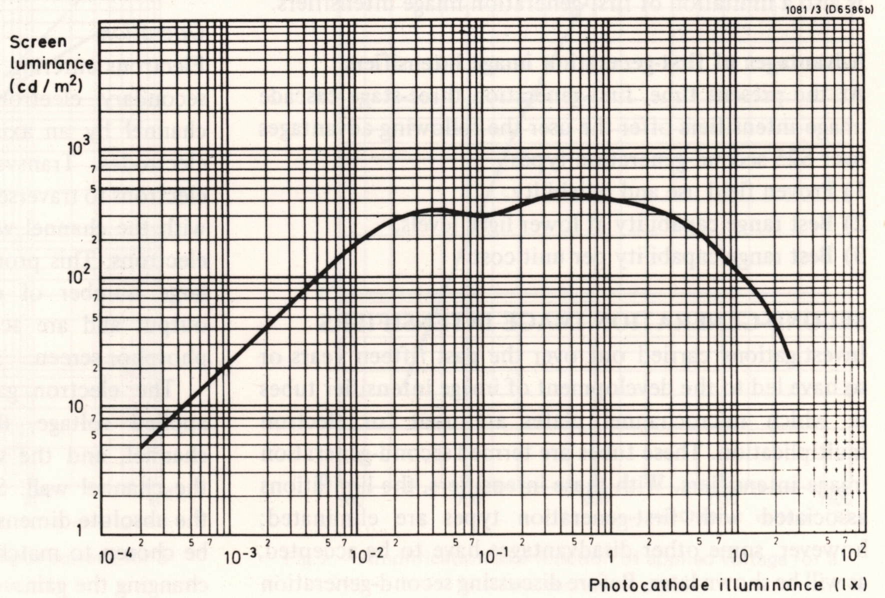


Fig.3 — Typical a.b.c. characteristic for cascade intensifier

Limitations of first-generation image intensifiers

In any image intensifier system, the total amplification needs to be sufficiently high to ensure that the signal-to-noise ratio at the retina of the eye for a detail in the scene is limited only by fluctuations in the impinging photons (or, more exactly, of the photoelectrons in the first stage). Subsequent links in the chain that transfers the image information should not degrade this ratio. In particular, the noise contribution of subsequent stages and that of the eye should not exceed the amplified photon noise. It has been found that to satisfy this requirement in a first-generation intensifier, the gain must be at least 50 000. In many applications, for example in aiming sights with relatively small image angles, it is not practicable to reduce the tube magnification, and therefore this level of intensification can only be achieved by employing three amplifying stages. This imposes a limitation on first-generation image intensifiers, namely one of physical volume and weight.

In general, the gain limit of 50 000 will not always be absolutely necessary; it depends to some extent on the purpose of observation. For aiming sights, resolution of the smallest detail is essential, and investigations in the laboratory and in the field have shown that for this application, high light amplification is more important than in, for example, night viewers for vehicle drivers.

With a three-stage cascade intensifier, the persistence of the three phosphor screens in series can be disturbing when observing scenes with moving highlights: bright tracks streak the image. Although the persistence of a type P20 phosphor is sufficiently short (a few milliseconds) for a single-stage tube, in multi-stage tubes the afterglow remains visible for several seconds as a result of the amplification. Phosphors with a much shorter persistence are available, but they are unsuitable because of their poor luminous efficiency. Thus image persistence is also a limitation of first-generation image intensifiers.

Advantages of first-generation image intensifiers

At the present time, first-generation three-stage cascade image intensifiers offer the user the following advantages over S25 second-generation types:

1) Proven field use and reliability,
2) Best range capability at lower light levels,
3) Best range capability per unit cost.

SECOND-GENERATION IMAGE INTENSIFIERS

Investigations carried out over the past fifteen years or so have led to the development of image intensifier tubes in which microchannel plates are used for electron multiplication. These tubes are termed second-generation image intensifiers. With these intensifiers, the limitations associated with first-generation types are eliminated; however, some other disadvantages have to be accepted, as will be shown later. Before discussing second-generation intensifiers, a brief description of the operation and characteristics of the microchannel plate may be found useful.

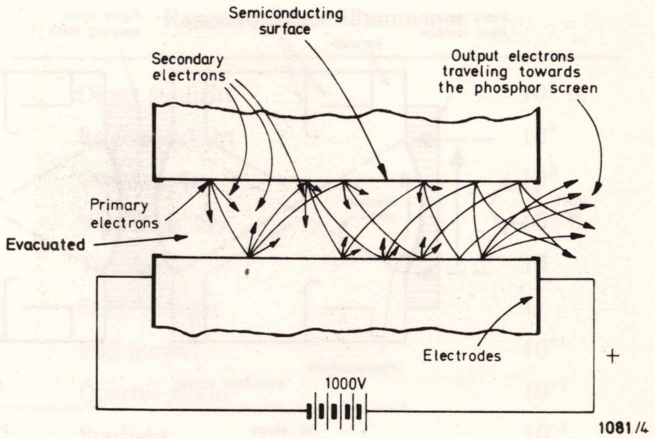


Fig.4 — Electron multiplication in one channel of a microchannel plate

Microchannel plates

A microchannel plate is an array of minute channel electron multipliers that can be used for amplifying an electron beam containing spatial information. Each cylindrical channel (see Fig.4) in the array combines the function of the dynode structure of a conventional photomultiplier with that of the resistor chain that distributes the supply voltage to the separate dynodes.

Each channel is about 40 times as long as its diameter. The inner surface is made slightly conducting by a reduction process, and will emit secondary electrons when bombarded with primary electrons that have been accelerated by an electric field. The resistance between electrodes deposited on each side of the microchannel plate, the plate resistance, is approximately $3 \times 10^8 \Omega$.

Operation

Electrons entering a channel strike the wall and produce secondary electrons which are accelerated along the channel by an axial electric field applied between the electrodes. Transverse energy of emission causes the electrons to traverse the channel so that they again collide with the channel wall and produce still more secondary electrons. This process is repeated many times so that a large number of electrons emerge from the channel output and are accelerated across a narrow gap to a phosphor screen.

The electron gain of a channel depends upon the applied voltage, the length-to-diameter ratio of the channel, and the secondary emission characteristics of the channel wall. Since the gain does not depend upon the absolute dimensions of the channel, the diameter can be chosen to match the resolution requirements without changing the gain.

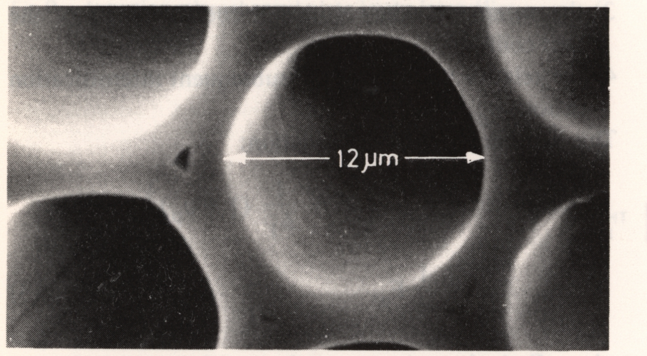


Fig.5 — Scanning electron micrograph of an etched face of a microchannel plate

Microchannel plates (see Fig.5) are currently manufactured with a centre-to-centre spacing of the individual channels of less than 15µm. In theory, it is possible to reduce this still further and hence achieve better resolution; however, such a reduction creates technological problems which have yet to be solved.

Electrical properties

The electrical requirements for the microchannel plate depend mainly on the specification of the image intensifier. On the one hand, the multiplied electron current must be sufficiently high to produce a bright image at the luminescent screen; on the other, a high output current implies a low microchannel plate resistance, which may result in an objectionably high power consumption. A typical output current/input current characteristic for a microchannel plate is shown in Fig.6. The relationship is approximately linear until the output current approaches the standing current, that is the current that flows continuously in the channel walls between the input and output electrodes. For a microchannel plate of resistance 100MΩ with 1kV across it, this standing current is about 10µA.

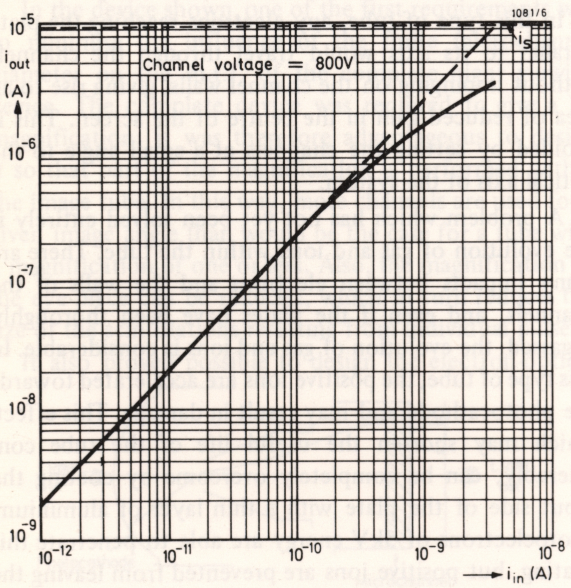


Fig.6 — Output current as a function of input current for a microchannel plate

The secondary electron current cannot appreciably exceed the standing current because the resulting electrical charging of the channel walls would then limit the secondary emission. In practice, linearity between input and output current is ensured if the output current does not exceed 5 to 10% of the standing current. If, however, a highlight does occur on a small area of the plate, localised saturation takes place and prevents excessive localised current densities.

Amplification in a microchannel plate is a rapid function of applied voltage (see Fig.7). Thus, amplification can be controlled over a wide range by varying the applied voltage. The gain of a cascade intensifier can be controlled in a similar manner by varying the oscillator voltage. However, only microchannel plate intensifiers operate with the inherent localised highlight control mentioned above. In a cascade tube, with a picture of low average brightness, a highlight will produce flare because the device is operating locally at very high gain. With a microchannel plate intensifier, however, it is possible to achieve complete protection against excessive highlights.

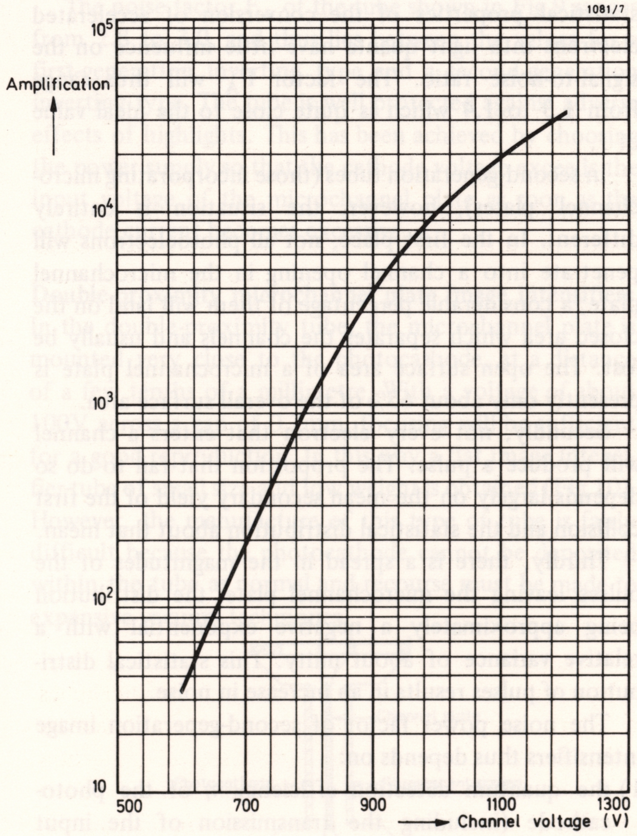


Fig.7 — Amplification as a function of applied voltage for a microchannel plate

Noise properties

In an image-intensifying system, the signal-to-noise ratio at the output will be inferior to that at the input. This deterioration in the quality of reproduction can be expressed in terms of a noise power factor F; that is:

$$F = \sigma_{in}^2/\sigma_{out}^2, \qquad \ldots(1)$$

where σ_{in} and σ_{out} are the signal-to-noise ratios at the input and output respectively. The signal-to-noise ratio σ_{in} at the input is determined by statistical fluctuations in the number of impinging photons, referred to a defined surface and a defined time unit.

The first deterioration in the signal-to-noise ratio occurs at the photocathode. If the quantum efficiency of the photocathode is η, then:

$$F = F_A/\eta, \qquad \ldots(2)$$

where F_A is the noise power factor of the amplifier stages after the photocathode. As a rule, η barely exceeds 10%, even at the point of maximum quantum efficiency.

It should be recognised that the factor F_A actually characterises the important differences between image intensifier tubes of the first and second generations. In first-generation tubes, electrons are lost only as a result of reflections at the luminescent screen. Also, the statistical properties of the conversion of accelerated electrons into light quanta have little influence on the signal-to-noise ratio. The factor F_A will thus range from 1.3 to 1.4 which is quite close to the ideal value of 1.

In second-generation tubes (those incorporating microchannel plates), however, the situation is entirely different. In the first place, not all photoelectrons will penetrate into a channel opening in the microchannel plate; a considerable percentage of them will land on the closed area which separates the channels and usually be lost. The open surface area of a microchannel plate is presently only about 65% of the overall surface area.

Secondly, not every electron that enters a channel will produce a pulse. The proportion that fail to do so depends largely on the mean secondary yield of the first collision and the statistical distribution about that mean.

Thirdly, there is a spread in the magnitudes of the pulses leaving the microchannel plate, the distribution being approximately a negative exponential with a relative variance of about unity. This statistical distribution of pulses results in an increase in noise.

The noise power factor of second-generation image intensifiers thus depends on:

1) the quantum detection efficiency η of the photocathode (including the transmission of the input window) for the incident light;
2) the detection efficiency D of the channel plate;
3) the mean secondary emission coefficient δ_1 of the first collision;
4) the relative variance of the subsequent intensification process υ_p.

This is summarised in the equation:

$$F = \frac{1}{\eta D}\left(1 + \frac{1}{\delta_1}(1 + \upsilon_p)\right),$$

$$= \frac{1}{\eta} F_a, \qquad \ldots(3)$$

where F_a is the noise power factor of the microchannel plate.

It should be noted that the visual impression of the noise in tubes of the first generation differs from that in tubes of the second generation. For first-generation types, the energy distribution of the light pulses is fairly narrow, whereas for second-generation types it has an exponential character. As a result, the few high-intensity pulses are apt to be particularly annoying. Investigations in the laboratory and in the field indicate that this effect can be greatly reduced, without affecting perception, by moderately increasing the amplification.

Inverting microchannel plate image intensifiers

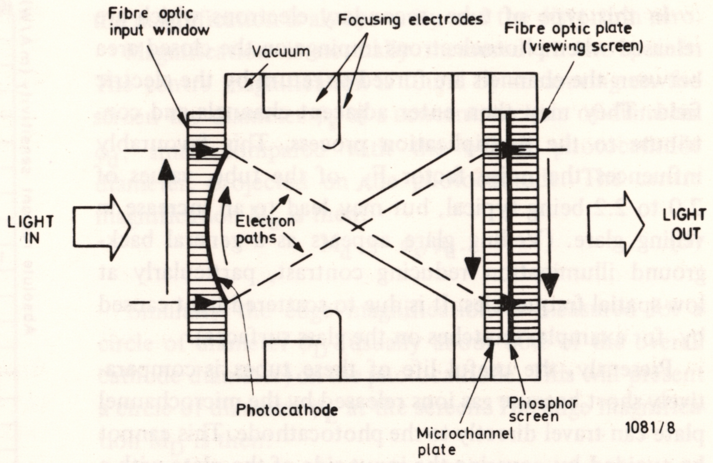
Tubes of this type are derived from the familiar first-generation electrostatically-focused intensifier (see Fig.1). The microchannel plate is mounted just in front of the luminescent screen (see Fig.8). The photoelectrons impinge on the microchannel plate and, after multiplication in the channels, are accelerated towards the screen which is mounted as close as possible to the plate. A field strength of about 5kV/mm across the gap between microchannel plate and screen gives a satisfactory modulation transfer function (m.t.f., see later).

In this type of tube, any electrons reaching the plate parallel to its axis would travel through the channels without impinging on the channel walls, giving rise to an area of reduced gain in the centre of the screen. This is avoided by setting the channels at a small angle to the optical axis of the system.

A problem which has not yet been solved entirely is the evolution of gas and ions within the tube. There are many impacts between electrons and the walls of the channels, and even if the plates have been thoroughly degassed, the evolution of gas and ions is considerable. In this type of tube, the positive ions are accelerated towards the photocathode and may result in damage. This effect, which may shorten the useful life of the tube considerably, can be completely overcome by coating the input side of the plate with a thin layer of aluminium. Photoelectrons of 5kV energy are able to penetrate this coating, but positive ions are prevented from leaving the channels in the direction of the photocathode.

Unfortunately, the aluminium coating increases the noise factor because it tends to disperse the photoelectrons. In practice, the noise factor F_a will assume a value ranging between 3 and 5. On the other hand, the aluminium coating does have the advantage that it prevents photoelectrons which are reflected from the closed area between the channels from entering a channel at another place, which would result in loss of contrast; the m.t.f. at low frequencies is thus preserved. Thus, the main features of inverting microchannel plate image intensifiers are their good life expectancy and good image contrast, particularly at low frequencies.

Fig.8 — Inverting microchannel plate image intensifier



Normally, the electron-optical lenses of intensifiers of this type are designed to achieve unity magnification of the image from input to output. However, to meet certain operational requirements, both magnifying and demagnifying intensifiers have been developed. Fig.9 shows a recently developed magnifying tube in which measures are taken to avoid some of the limitations discussed above. Its electrode configuration is represented only schematically in Fig.9; its operation is as follows.

In the device shown, one of the first requirements was to minimise the influence of the finite microchannel diameter on the m.t.f. of the complete night viewing device. The complete device was required to give a ×5 magnification; it was therefore advantageous to design it so that part of the magnification is performed within the image tube. In this way, more channels are used for a given image angle than would be the case for a tube with a magnification of one or less. Also, the magnification of the eye-piece can be reduced, which in turn reduces the size of the scintillations resulting from individual pulses.

It also proved possible to design the electron-optical configuration so that positive ions produced at the microchannel plate are prevented from reaching the photocathode. The aluminium coating of the microchannel plate could thus be dispensed with, which allows the primary electron energy to be reduced to about 1kV, thereby improving the secondary emission. At the same time, the image distortion was reduced to 2%. In this tube, secondary electrons produced in the closed area between the channels are not recaptured, as in the case of the double-proximity tube (see later), but are collected by the positively-biased anode. They are thus prevented from reducing the low-frequency contrast.

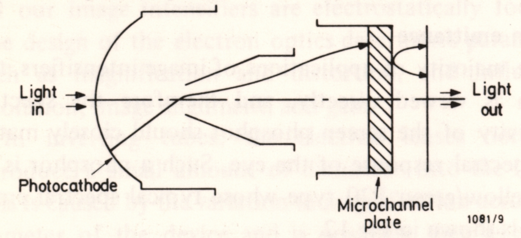


Fig.9 — Magnifying microchannel plate image intensifier

The noise factor F_A of the tube shown in Fig.9 ranges from 2.8 to 3.0, and thus lies between the values for a first-generation inverting type and a second-generation inverting type. The tube is well protected against adverse effects of highlights. This has been achieved by choosing the power supply so that the cathode voltage exceeds the input voltage of the microchannel plate as soon as the cathode current becomes excessive.

Double-proximity microchannel plate image intensifiers

In the double-proximity tube, the microchannel plate is mounted very close to the photocathode, at a distance of a few tenths of a millimetre. With a voltage of about 100V across a gap of 0.2mm, focusing will be sufficient for a good reproduction. In this way a flat image intensifier tube of small size and low weight is obtained (Fig.10). However, the manufacture of this type of tube is fairly difficult because the photocathode cannot be deposited within the tube as normal and recourse must be made to expensive vacuum techniques.

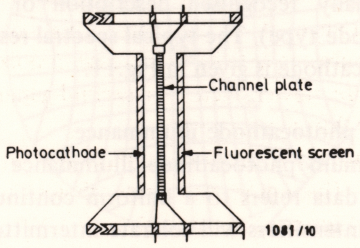


Fig.10 — Double-proximity microchannel plate image intensifier

In this type of tube, secondary electrons which are released when photoelectrons impinge on the closed area between the channels are forced to return by the electric field. They may then enter adjacent channels and contribute to the multiplication process. This favourably influences the noise factor F_A of the tube, values of 2.0 to 2.2 being typical, but may lead to an increase in veiling glare. (Veiling glare appears as a general background illumination reducing contrast, particularly at low spatial frequencies. It is due to scattered light caused by, for example, scratches on the glass surfaces).

Presently, the useful life of these tubes is comparatively short because gas ions released by the microchannel plate can travel directly to the photocathode. This cannot be avoided by covering the input side of the plate with a thin aluminium coating because this would form a barrier to the low-energy photoelectrons. Several attempts have been made recently to solve this problem; for example, by the application of a very thin film of aluminium oxide, or by the use of curved channels. Both these measures reduce the ion current towards the photocathode without unduly impairing the performance. Double-proximity tubes are mainly used in those applications where small size and weight are of prime importance, for example in night-viewing goggles.

CHARACTERISTICS OF IMAGE INTENSIFIERS AND THEIR MEASUREMENT

The main characteristics of image intensifier tubes are described below. Resolution and modulation transfer function are discussed later, under a separate heading.

Photocathode sensitivity

The properties of the photocathode are described in terms of its spectral response and sensitivity. The sensitivity can be expressed in two ways: luminous sensitivity (μA/lm) and radiant sensitivity at a specified wavelength (mA/W). The luminous sensitivity is measured with a photometer and a tungsten filament lamp operating at a nominal colour temperature of 2856K. To enable radiant sensitivity to be measured, the lamp is filtered to give wavelengths of 800 and 850nm.

Passive night-vision applications require photocathodes with high luminous and radiant sensitivities. The S25 photocathode is ideal for such applications and offers excellent response to night-sky illumination (S25 is an internationally recognised description of a particular photocathode type). The typical spectral response of the S25 photocathode is given in Fig.11.

Maximum photocathode illuminance

The maximum photocathode illuminance given in the published data refers to a uniform continuous illuminance. All intensifiers will tolerate intermittent bursts of cathode illuminance which are much higher than the rated maximum value. However, prolonged exposure to any source of bright illumination will shorten the life of the tube. Most intensifiers incorporate automatic brightness or gain controls (a.b.c. or a.g.c.). These reduce the screen luminance but do not reduce the photocathode current. Whenever possible the photocathode illuminance should comply with the recommended operating conditions.

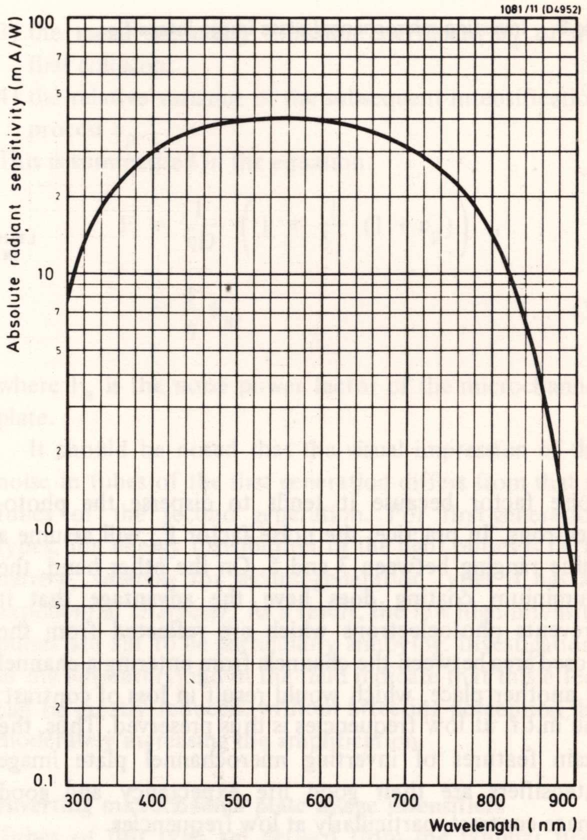


Fig.11 — Typical spectral response of S25 photocathode

Mean screen luminance

This is the luminous intensity (cd) of the screen averaged over a specified area (m^2). It is given only for intensifiers with an integral power supply, and is a function of the properties of the power supply.

Screen luminance ratio

This is the ratio of the luminance at the centre of the screen and the average luminance at any point on a defined concentric circle.

Screen emittance

In the majority of applications of image intensifiers, the screen is viewed directly, and therefore the spectral emissivity of the screen phosphor should closely match the spectral response of the eye. Such a phosphor is of the yellow/green P20 type whose typical spectral emittance is shown in Fig.12.

The decay time or persistance of the phosphors used in intensifier tubes varies with the level of excitation by the impinging electrons. It is defined as the time taken after the removal of excitation for the screen luminance to fall to 36% of its initial peak value, and is normally in the range 0.5 to 5ms. For cascade image intensifiers, there are three phosphors in series; hence, the persistance of the output phosphor will appear much longer than that of a single phosphor.

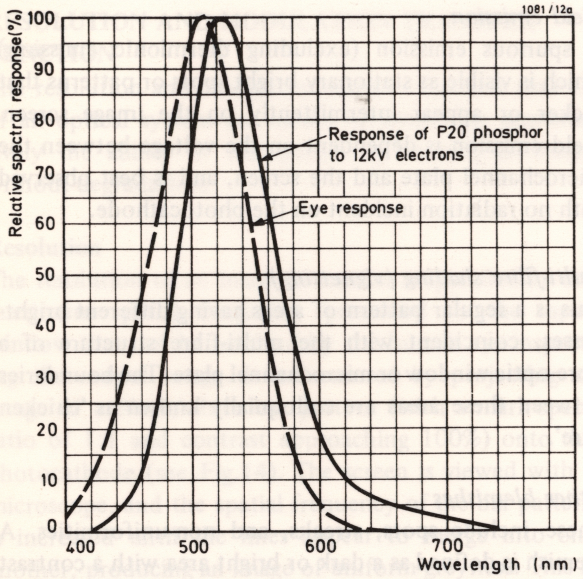


Fig.12 — Typical spectral emittance of P20 phosphor and spectral response of dark-adapted eye

Gain

The gain of an image tube can be defined as:

$$G = \pi L_o / E_i, \qquad \ldots (4)$$

where L_o is the luminance (cd/m^2) in a direction normal to the screen, measured with an eye-corrected photometer, and E_i is the uniform illuminance (lx) incident on the entire area of the photocathode. The illuminance is produced by a tungsten lamp at a nominal colour temperature of 2856K.

Gain is dimensionless. It can be expressed alternatively as the output luminance divided by the input illuminance; that is, in nits/lx. The resulting value is a factor of π smaller than the dimensionless value.

Magnification and distortion

All our image intensifiers are electrostatically focused. The design of the electron optics determines parameters such as magnification and distortion, and influences resolution, image alignment, and gain.

In inverting tubes, the electron lenses normally introduce a small amount of distortion into the image. This is caused by the variation in magnification across the diameter of the device and is normally seen as a pincushion effect (see Fig.13). In double-proximity tubes, the magnification is always unity and the distortion zero.

Magnification is normally measured at two points. The centre magnification is found by measuring on the screen the diameter ϕ_s of a concentric circle of diameter ϕ_d, small compared with the overall photocathode diameter, projected on the photocathode. The centre magnification M_d is then:

$$M_d = \phi_s / \phi_d . \qquad \ldots (5)$$

Similarly the edge magnification is measured for a circle of diameter ϕ_D (usually about 80% of the overall cathode diameter) on the photocathode. This will present a circle of diameter ϕ_E at the screen. The edge magnification M_D is then:

$$M_D = \phi_E / \phi_D . \qquad \ldots (6)$$

Because of the difficulty in measuring small differences in the diameter ϕ_s, there can be significant variations in the value of M_d unless very careful precautions are taken.

Distortion in image intensifiers is expressed as:

$$\%\,\text{distortion} = \left(\frac{M_D}{M_d} - 1 \right) \times 100. \qquad \ldots (7)$$

Image alignment

The geometrical and optical axes of the intensifier may not coincide. Image alignment is a measure of this. It is the distance on the screen between the geometric axis and the image of a point at the geometric centre of the photocathode.

Background noise

Noise is expressed in terms of background equivalent illuminance (B.E.I.), also known as equivalent background input (E.B.I.), which is the input illuminance required to give an increase in screen brightness equivalent to the background brightness. With the supply voltage applied and no input illumination incident on the photocathode, the screen will have a finite background

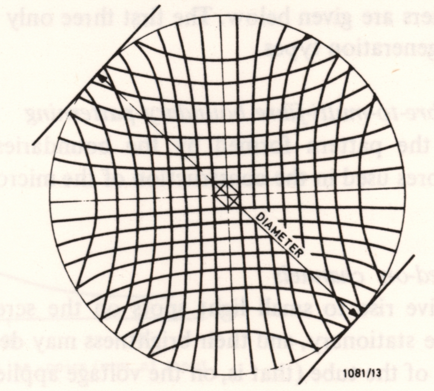


Fig.13 — Pincushion effect

brightness, which may be caused by one or more of the following effects.

Thermionic emission of the photocathode
This is dependent on the cathode temperature.

Electron scintillations
These are usually observed only in high-gain tubes by a dark-adapted observer using a magnifier, and without photocathode illumination. They are seen as small bright spots whose brightness and position vary with time, and they are caused by individual electrons emitted by the photocathode. The spots appear as a uniform scintillating field over the whole area of the screen, and are too numerous to count. They account for the greater part of the background brightness when the photocathode is not illuminated.

Ion scintillations
These are very bright spots whose brightness and position vary with time but which are only seen near the centre of the screen. They appear bigger than electron scintillations. They can be seen with the unaided eye (depending on the gain of the tube under observation) but are best observed with a magnifier. They are caused by the emission of several electrons from the photocathode resulting from the impact of one ion. The density of ion scintillations is usually much lower than that of electron scintillations. Ion spots are best observed with some illumination incident on all areas of the photocathode except for the central area. The spots are then visible against a dark background.

Long-term phosphorescence of the screen
This may result from previous operation of the tube or from previous exposure of the screen to high levels of illumination of a spectral distribution which can excite phosphorescence. It is therefore recommended that image tubes are stored in darkness.

Picture quality
The main defects affecting picture quality in image intensifiers are given below. The first three only occur in second-generation types.

Multi-fibre-to-multi-fibre boundary patterning
This is the pattern formed by the boundaries of the multi-fibres used in the construction of the microchannel plate.

'Switched-on' channels
These give rise to small light spots on the screen. The spots are stationary, and their brightness may depend on the gain of the tube (that is, on the voltage applied across the microchannel plate).

Field emission
A spurious emission (excluding thermionic emission) which is visible as stationary bright spots or patterns that flicker or appear intermittently on the image screen. Field emission is dependent on the voltage between the microchannel plate and the screen, and is best observed with no radiation incident on the photocathode.

Multi-fibre shading (vignetting)
This is a regular pattern of areas having different brightnesses, coincident with the multi-fibre structure of a fibre-optic window or microchannel plate. The boundaries between these areas are colloquially known as 'chicken wire'.

Other blemishes
These include spots, streaks, and non-uniformities. A blemish is defined as a dark or bright area with a contrast greater than 30% with respect to the immediately surrounding area.

The picture quality of all intensifiers is assessed using a magnifier of approximately X5 magnification. Note that, because of the transfer characteristics of typical tv systems, there is a significant difference in the subjective appearance of an image tube screen viewed directly compared with one viewed through a tv system.

Recovery time
Integral-oscillator cascade image intensifiers and microchannel plate image intensifiers are designed to recover rapidly after fast changes in illuminance. The recovery time given in the published data is the time taken for a useful image to be restored on the screen after the photocathode illuminance is changed rapidly from the maximum rated value to near zero, and *vice-versa.*

Environmental tests
The environmental requirements included in the target specifications are determined by the application at which the device is aimed. Most image intensifiers are required to pass several environmental tests. The tests are:

1) high-temperature and low-temperature storage, operation, and thermal shock tests;
2) humidity tests;
3) mechanical shock and vibration tests.

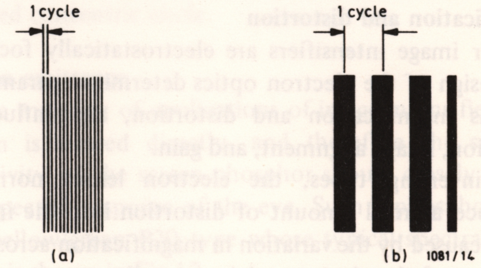


Fig.14 — Illustration of the meaning of spatial frequency:
(a) high frequency (b) low frequency

RESOLUTION AND MODULATION TRANSFER FUNCTION

The resolution and modulation transfer function (m.t.f.) of an optical system are means of describing quantitatively the ability of the system to transmit an image without degradation.

Resolution
The resolution of an image intensifier is specified by the maximum spatial frequency presented to the input window that can be resolved by the human eye inspecting the output image. It is measured by projecting an image of a black/white bar pattern (with a mark-space ratio of 1:1 and contrast approaching 100%) onto the photocathode (see Fig.14). The screen is viewed with a microscope, and the spatial frequency of the bar pattern is increased until the lines appear to merge into one another, producing an image of uniform greyness. This is the limit of resolution. Resolution is usually measured at two points: at the centre of the image, and at a point on a circle of diameter $\phi_E = 0.7\phi_D$.

Modulation transfer function
The m.t.f. of an optical system is best specified by a graph (or graphs) such as Fig.15; the published data gives values for three spatial frequencies.

The m.t.f. is measured by presenting the system with an image that contains either a wide range of spatial frequencies, from low to high values, or one in which the spatial frequency is slowly varied from low to high values. In the second case, it is a comparatively easy matter to plot a graph of modulation (output image contrast ratio, assuming an input contrast ratio of 100%) against spatial frequency. Analysis of the data, from the first mentioned measurement method, requires the data to be Fourier transformed – possibly using a dedicated computer.

A crude analogy can be drawn between the concepts of resolution and cut-off frequency, and between m.t.f. and frequency response.

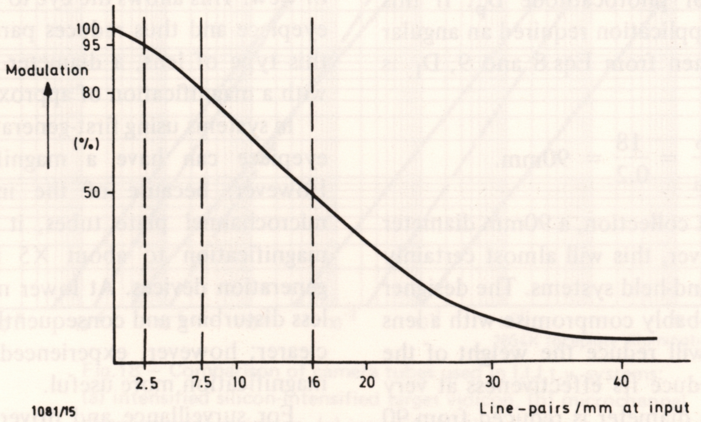


Fig.15 – Typical m.t.f. curve

IMAGE INTENSIFIER SYSTEMS

Image intensifier systems fall into two categories: direct view systems, and low-light-level television systems (l.l.l.t.v.). The former are devices that can be viewed directly by eye; they are mainly used for night viewing applications such as surveillance, aiming sights, and driver viewers. L.L.L.T.V. systems are normally used for remote surveillance, although other applications such as night flying of helicopters and the examination of aero engines using fibre scopes are being examined.

Direct view systems
Fig.16 shows a schematic diagram of a direct view system. There are three basic components: objective lens, image intensifier, and eyepiece. Of these, the most important, and often the most neglected, item is the objective lens.

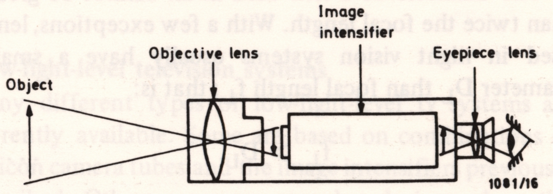


Fig.16 – Schematic diagram of direct view system

This lens collects the light which is then intensified by the image tube. Unless the lens collects sufficient light, the system will not be usable, no matter how effective the image intensifier.

Objective lens
The choice of objective lens depends on the application for which the image intensifier is intended. For aiming applications and long-range surveillance, the angular field of view of the system is small (see Fig.17a), allowing the use of small image tubes; for driver viewers and short-range surveillance, however, the angular field of view is required to be large (see Fig.17b) and consequently an image tube with a large-diameter photocathode is necessary. For example, for aiming sights the field of

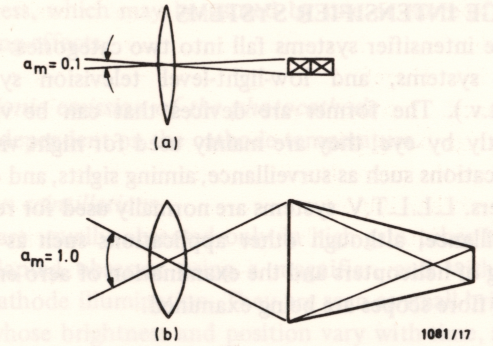


Fig.17 — Effect of angular field of view on size of input window

view α_m is about 0.1rad, whilst for driver viewers it is about 1rad.

The diameter of a lens used in air cannot be greater than twice the focal length. With a few exceptions, lenses used in night vision systems usually have a smaller diameter D_L than focal length f_L; that is:

$$f_L \geqslant D_L. \qquad \ldots(8)$$

From the focal length and the angular field of view α_m required, it is possible to calculate the optimum diameter D_p of the photocathode; thus:

$$D_p = f_L\,\alpha_m \geqslant D_L\,\alpha_m. \qquad \ldots(9)$$

In practice, for a small angular field of view, D_p will be determined by the maximum tolerable dimensions of the objective system. In applications requiring a large angular field of view, the limiting factor is D_p itself.

Because the diameter D_L of the objective lens limits the amount of light detected by the system, it is important that D_L be made as large as possible. The size of D_L will depend on: 1) the angular field of view, 2) the maximum tolerable length of system, and 3) the maximum tolerable weight. Items 2) and 3) may well limit the choice of image intensifier which will in turn decide the maximum diameter of photocathode D_p. If this were, say, 18mm and the application required an angular field of view of 0.2rad, then from Eqs.8 and 9, D_L is given by:

$$D_L \leqslant f_L = \frac{D_p}{\alpha_m} = \frac{18}{0.2} = 90\text{mm}.$$

Thus for maximum light collection, a 90mm diameter lens should be used. However, this will almost certainly be too heavy for use in hand-held systems. The designer in such cases will then probably compromise with a lens of 50mm diameter. This will reduce the weight of the system, but it will also reduce its effectiveness at very low light levels. If the lens diameter is reduced from 90 to 50mm, the light collected is reduced in proportion to the loss in area; thus:

$$E_{50} = \frac{(50)^2}{(90)^2} \cdot E_{90} = 0.31\,E_{90},$$

where E_{50} and E_{90} are the amounts of light collected with 50mm and 90mm lenses respectively. Therefore, by limiting the weight of the system (or equally, limiting the maximum lens diameter), the designer has reduced the light collected to less than one-third of the optimum. Thus, if the optimum system was capable of detecting the target reflecting 100μlx of light, the compromised system will only detect the same target if the light level is increased to 300μlx, or, if the target is brought nearer to the objective lens by a factor r, where r^2 is 0.31. Therefore, the range at which the target can be detected is reduced to 55% of the optimum.

This example shows, quite dramatically, how important the lens diameter is in the design of a night viewing system. It is quite clear that whenever possible, low weight and small size should be sacrificed for an increased lens diameter. It is obvious that no matter how effective an image intensifier is, it can never perform to its best unless the objective lens allows it to do so.

It is therefore important to use lenses that have the largest possible physical diameters. However, it should be remembered that lenses with small f-numbers (large aperture) frequently have inferior modulation transfer characteristics, particularly off-axis. Thus, unless great care is taken with the design of the lens, the loss from reduced contrast in the image can more than offset the gain from efficient light-gathering. In order to obtain the optimum contrast performance from the intensifier, the objective lens should have high m.t.f. at all spatial frequencies up to 30 line-pairs/mm and should not be less than 75% at 10 line-pairs/mm.

Eyepiece

The design of the eyepiece is critical to the subjective appearance of the image. For aiming sights, it is normal to choose an eyepiece which presents a small output field of view. This allows the eye to be brought up close to the eyepiece and thus reduces parallax errors in aiming. For this type of lens, a diameter of 7 to 10mm is normal with a magnification of approximately X7.

In systems using first-generation image intensifiers, the eyepiece can have a magnification more than X10. However, because of the inherent noise in inverter microchannel plate tubes, it is advisable to limit the magnification to about X5 in systems using second-generation devices. At lower magnifications, the noise is less disturbing and consequently the information appears clearer; however, experienced observers will find X10 magnification more useful.

For surveillance and driver viewer applications, it is more comfortable to use both eyes to view the screen. This can be accomplished using either two separate systems, one for each eye, or a single lens eyepiece for viewing by both eyes. Consequently, many systems utilise binocular eyepieces. These have large diameters and normally allow the observer to view the screen from a distance of 15 to 45cm. A magnification of X5 is the maximum that can be used satisfactorily.

System performance

The importance of the objective lens has already been discussed. It is shown again when an attempt is made to calculate the brightness of the output image of the intensifier. To make this calculation, the following assumptions are made:

1) the objective lens is focused at infinity,
2) the target is large enough to be significant.

Then, the luminance of the output screen of the intensifier is given by:

$$L = \frac{IRTG}{4f^2}\ \text{cd/m}^2, \qquad \ldots(10)$$

where L is the luminance of the output screen, I is the illuminance of the target, R is the reflectivity of the target, T is the transmission of the objective lens, G is the gain (including a factor for the photocathode sensitivity), and f is the effective f-number of the objective lens.

Thus, for an f/1.47 lens having an 80% transmission, the screen luminance from a scene of 10% reflectivity illuminated by 10mlx is 4.6cd/m².

Note that at very low light levels, the contribution of the photocathode sensitivity to the overall gain becomes significant and is important when considering signal-to-noise ratio.

When the scene contains highlights or bright flashes, the screen luminance of a cascade image intensifier may well become so high that the contrast of the image is degraded. Therefore, in applications where this is likely to occur, second-generation tubes should be used, advantage being taken of the point highlight saturation effect of the microchannel plate (see earlier).

The other important system parameter is its overall magnification. The dimensions of the objective and eyepiece lenses having been set, the system magnification is given by:

$$M = \frac{M_l f_o}{f_e},$$

where M is the angular magnification, M_l is the linear magnification of the intensifier, f_o is the focal length of the objective lens, and f_e is the focal length of the eyepiece.

Low-light-level television systems

Many different types of low-light-level tv systems are currently available. Some are based on combinations of vidicon camera tubes and the image intensifiers previously described. Others are more complex devices where the intensifier and vidicon are contained in the same vacuum envelope. In the latter category, the intensified silicon vidicon (ISV) is the main device. This consists of an image intensifier, in series with a silicon-target vidicon, all in a single envelope. It is possible to increase the sensitivity of the ISV by fibre-optically coupling a single-stage first-generation image intensifier to the front of the triode intensifier. The device then becomes an intensified ISV, or I^2SV. The relative sensitivities of various l.l.l.t.v. devices are compared in Fig.18.

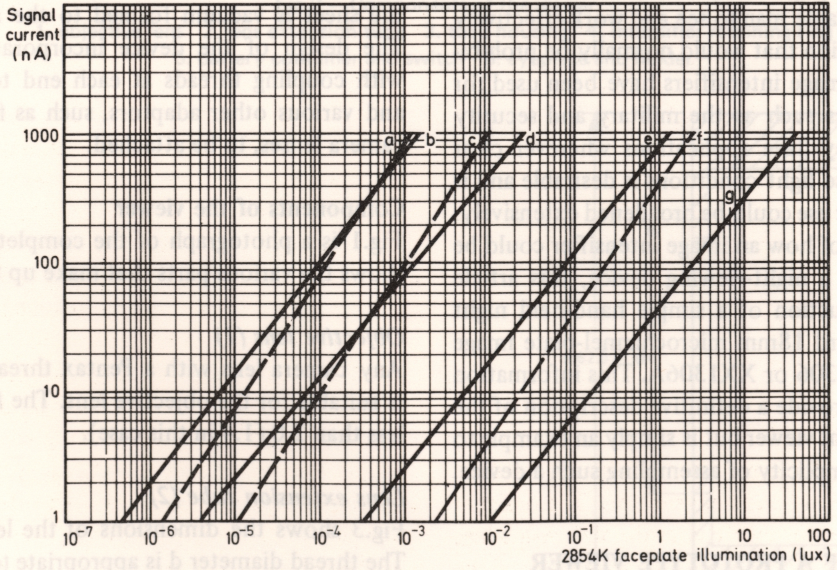


Fig.18 — Comparison of camera tubes used in l.l.l.t.v. systems: (a) intensified silicon-intensified target vidicon, (b) microchannel plate intensified vidicon, (c) intensified silicon vidicon, (d) two-stage cascade intensified vidicon, (e) single-stage intensified vidicon, (f) silicon vidicon, (g) vidicon

A simple hand-held night viewer

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This article describes the construction of a relatively inexpensive night-viewing device incorporating a Mullard image intensifier. Appropriate materials are suggested, and a possible method of assembly is outlined.

INTRODUCTION

Image intensifiers are electron-optical devices which have been developed to enable man to see and work effectively in levels of illumination that would normally be prohibitive. Traditionally, image intensifiers have been used for specialist applications such as the military and security fields. However, they have applications wherever clear viewing under adverse light conditions is desirable and it is envisaged that their use could be broadened extensively.

As an indication of how an image intensifier could be used as a basis for a night-viewing device, this article describes the construction of a simple hand-held night viewer using a Mullard 18mm microchannel-plate image intensifier type XX1306 or XX1306A. This information is not intended to provide a definitive description of the construction of a night viewer but is simply an example to show the ease and simplicity of assembling such a device.

CONSTRUCTION OF A PROTOTYPE VIEWER

A simple monocular device which can be hand-held is to be assembled, and the viewer is expected to be reasonably presentable and durable.

Construction materials

The overall requirements suggest that plastics are appropriate for the construction of the assembly. The tube housing and lens extension tube are machined from polyacetal which has a low specific gravity and good impact strength, and the threads can be cut with relative ease. The material can be moulded if production quantities are required. The eyepiece mount and tube body cover are made from acylonitrile butadiene styrene (ABS). This material can be thermally formed and is available in sheet-form with various surfaces, such as the 'leatherette' chosen. By making simple wooden formers, the sheet is vacuum formed to the appropriate shapes. The design of the device incorporates a tube module with coupling threads at each end to allow lens cones and various other adaptors, such as for standard 35mm camera lenses, to be attached.

Components of the viewer

Fig.1 is a photograph of the completed assembly. Fig.2 shows the various parts that make up the assembly.

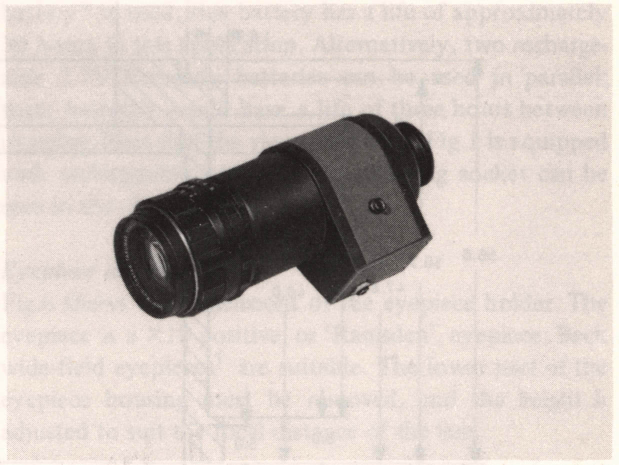
Objective lens (1)

Any camera lens with a Pentax thread or similar fitting is suitable for the objective lens. The f-number should be less than 2.0 (1.4 in this case).

Lens extension tube (2)

Fig.3 shows the dimensions of the lens extension tube. The thread diameter d is appropriate to the objective lens used, M42 × 1 for a Pentax lens, for instance. The distance L is the back focal length of the objective lens less the distance between the photocathode and the end

Fig.1 – A hand-held night viewer incorporating an image intensifier, Mullard type XX1306 or XX1306A



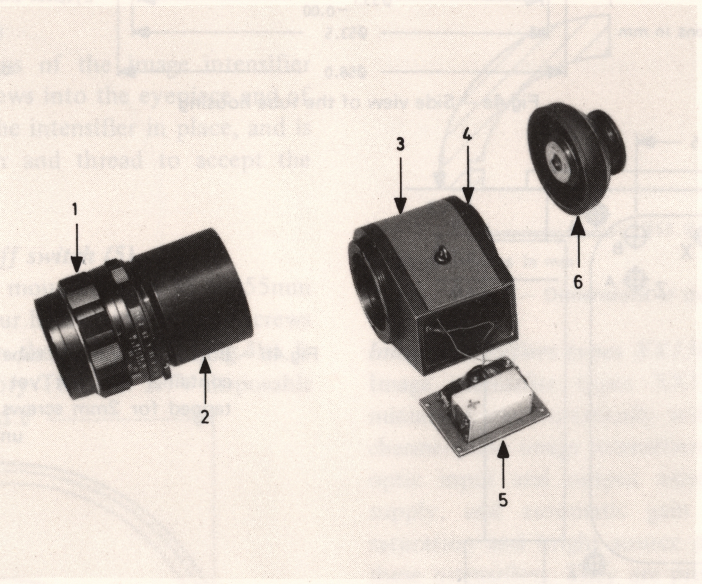


Fig.2 – Components of the viewer
1. objective lens 2. lens extension tube 3. tube housing 4. intensifier retaining ring
5. battery container and switch 6. eyepiece and holder

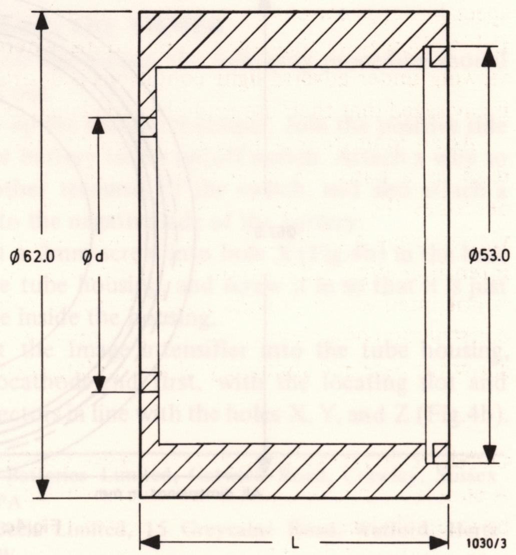


Fig.3 – Dimensions of the lens extension tube

All dimensions in mm

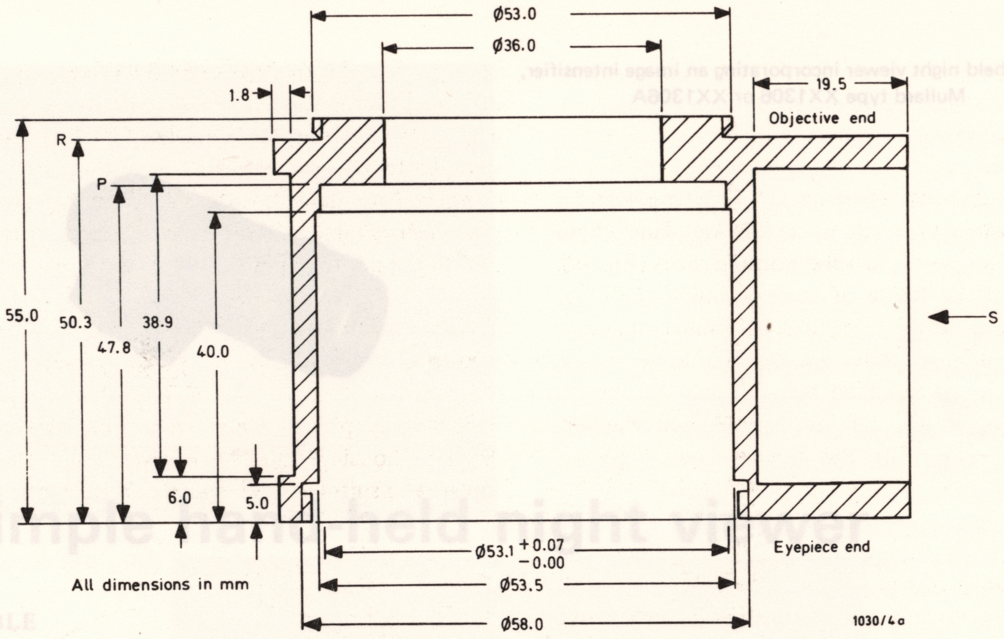


Fig.4a – Side view of the tube housing

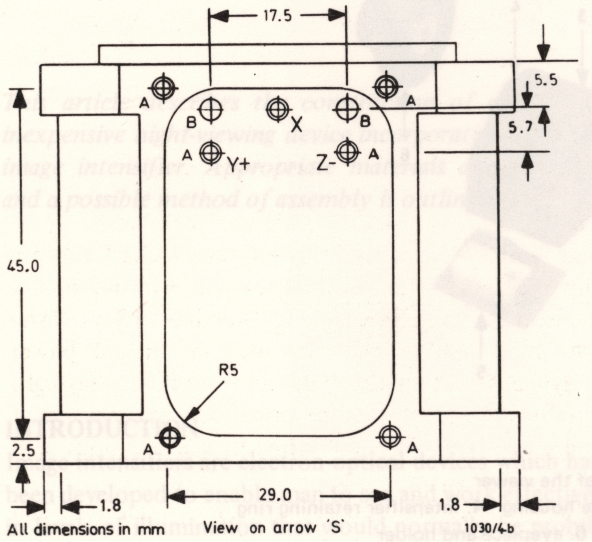


Fig.4b – Bottom view of the tube housing. Note that the battery container has not yet been attached. Holes 'A' are tapped for 2mm screws. Holes 'B' are 2mm diameter unthreaded

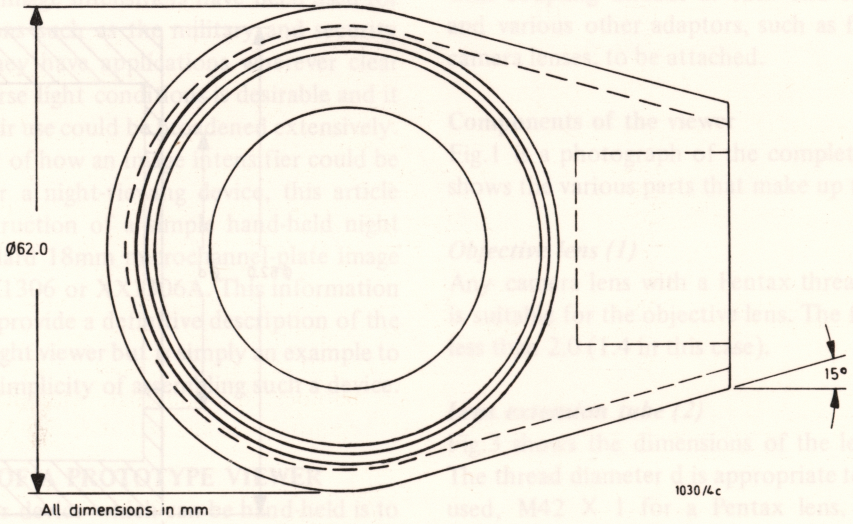


Fig.4c – Front view of the tube housing

of the tube holder. The back focal distance of the lens used is 42.5mm, giving L a value of 40mm.

Tube housing (3)

Fig.4 shows the dimensions of the tube housing. Plane P is the photocathode plane and plane R is the plane where the lens extension tube and tube housing meet (Fig.4a). The diameter of the inside of the housing is the most critical dimension. The image intensifier must fit inside the holder but do so tightly. Threaded hole X takes a locating screw, while threaded holes Y and Z (Fig.4b) take screws to which terminal tabs are attached which in turn make contact with the positive and negative terminals of the intensifier via pins inserted in the 'B' holes.

Intensifier retaining ring (4)

Fig.5 shows the dimensions of the image intensifier retaining ring. This ring screws into the eyepiece end of the tube housing to keep the intensifier in place, and is of an appropriate diameter and thread to accept the eyepiece holder.

Battery container and on/off switch (5)

The battery and switch are mounted on a 35 by 55mm piece of polyacetal with four holes to take 2mm screws to attach the assembly onto the tube holder (Fig.4b). It is suggested that one Mallory TR132N 2.7V disposable battery* is used; this battery has a life of approximately 30 hours in this application. Alternatively, two rechargeable 2.7V Eveready batteries can be used in parallel; these batteries would have a life of three hours between charging. Note that the viewer shown in Fig.1 is equipped with rechargeable batteries; the charging socket can be seen in the photograph.

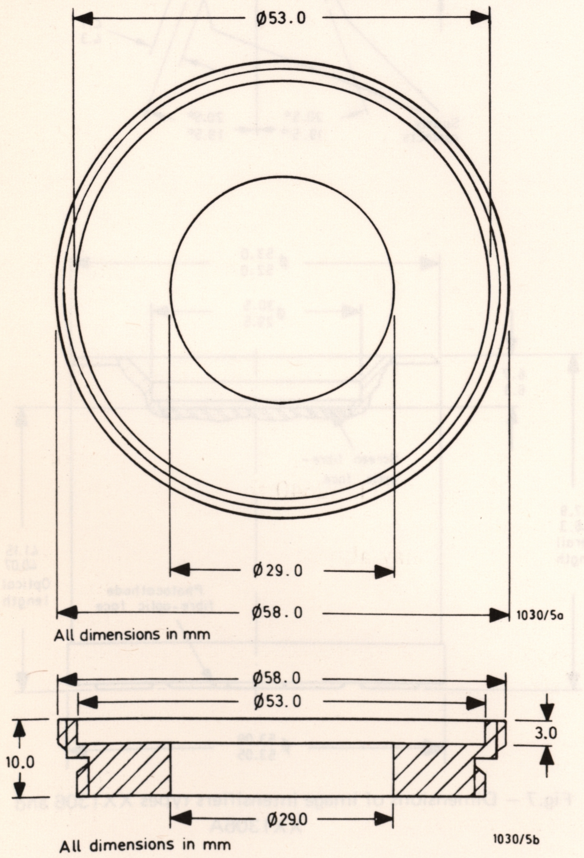


Fig.5 – Dimensions of the intensifier retaining ring

Eyepiece and holder (6)

Fig.6 shows the dimensions of the eyepiece holder. The eyepiece is a ×10 positive, or 'Ramsden', eyepiece; Beck wide-field eyepieces† are suitable. The lower part of the eyepiece housing must be removed, and the height h adjusted to suit the focal distance of the lens.

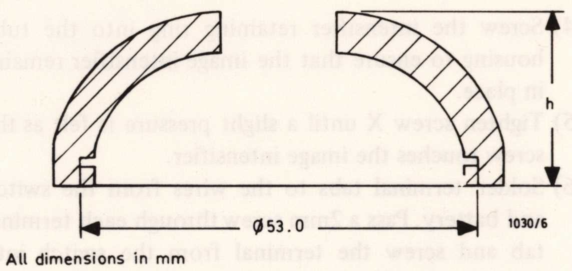


Fig.6 – Dimensions of the eyepiece holder

Image intensifiers types XX1306 and XX1306A

Image intensifier types XX1306 and XX1306A are miniature, electrostatically self-focused inverting microchannel-plate image intensifiers. They have 18mm fibreoptic input and output windows, an integral power supply, and automatic gain control. Point highlight saturation and bright source protection are features of these intensifiers. They are primarily intended for use in night-vision systems.

Table 1 gives abridged data for these intensifiers and Fig.7 shows their dimensions.

ASSEMBLING THE VIEWER

To assemble the viewer, the following procedure should be carried out.

1) Make up the battery container. Join the positive side of the battery to the on/off switch. Attach a wire to the other terminal of the switch, and also attach a wire to the negative side of the battery.
2) Insert a 2mm screw into hole X (Fig.4b) in the base of the tube housing, and screw it in so that it is just visible inside the housing.
3) Insert the image intensifier into the tube housing, photocathode-end first, with the locating slot and connectors in line with the holes X, Y, and Z (Fig.4b).

---

* Mallory Batteries Limited, Gatwick Road, Crawley, Sussex RH10 2PA

† Ealing Beck Limited, 15 Greycaine Road, Watford, Herts WD2 4PW

TABLE 1
Image Intensifier Data

Type No.	Photocathode			Screen		Resolution (line pairs/mm) typical	Gain limits	Operating voltage (V)	Mass (g) approx.
	Type	Diam. (mm)	Sensitivity (μA/lm) min.	Type	Diam. (mm)				
XX1306	25	18	200	P20	18	25	23000– 46000	2.7 d.c.	175
XX1306A	25	18	175	P20	18	25	10000– 50000	2.7 d.c.	175

4) Screw the intensifier retaining ring into the tube housing to ensure that the image intensifier remains in place.
5) Tighten screw X until a slight pressure is felt as the screw touches the image intensifier.
6) Solder terminal tabs to the wires from the switch and battery. Pass a 2mm screw through each terminal tab and screw the terminal from the switch into hole Y, ensuring that the other end of the terminal tab passes over a pin inserted into the appropriate 'B' hole (Fig.4b), thus forming a pressure contact with the connector on the image intensifier. Repeat the operation for the wire from the battery and a terminal tab screwed into hole Z.
7) Fix the battery container to the tube housing with four 2mm screws.
8) Screw the lens extension tube onto the housing, and then screw the lens onto the extension tube.
9) Screw the eyepiece holder onto the other end of the tube housing.
10) Reduce the aperture of the objective lens to a minimum. Turn on the intensifier in a darkened room and adjust the position of the eyepiece for optimum focus; minor adjustments to the distances L and h may be necessary

The viewer is now complete.

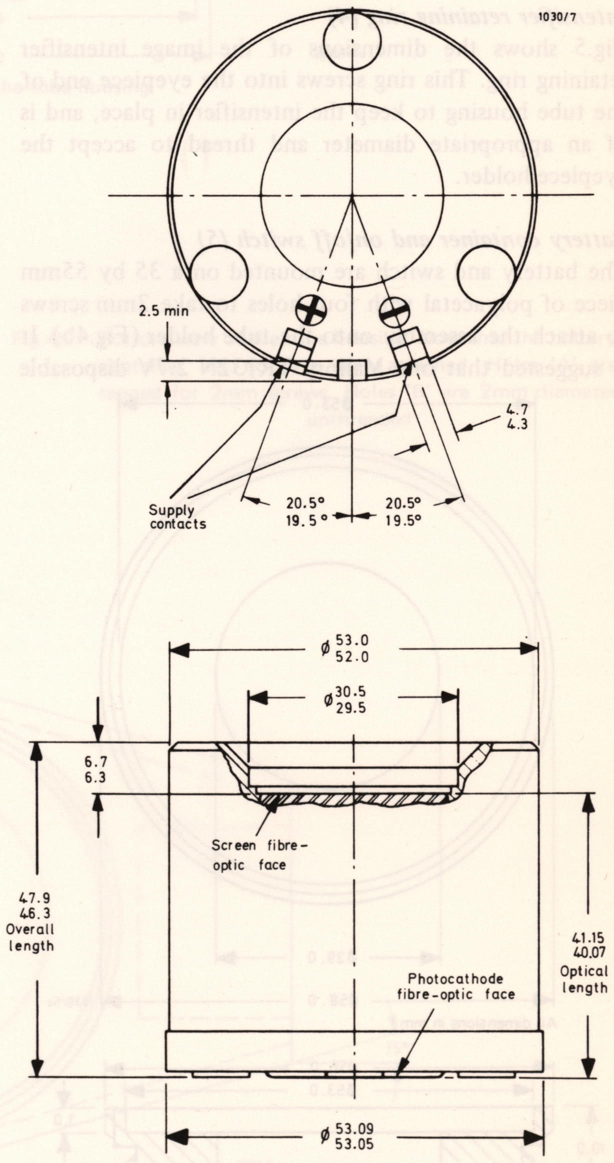


Fig.7 – Dimensions of image intensifiers types XX1306 and XX1306A

ACKNOWLEDGEMENT
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